1. (WO2002015433) TRANSMISSION DIVERSITY COMMUNICATION DEVICE

Latest bibliographic data on file with the International Bureau

Pub. No.: WO/2002/015433 **International Application No.:** PCT/JP2000/005380
Publication Date: 21.02.2002 **International Filing Date:** 10.08.2000
Chapter 2 Demand Filed: 26.12.2000

IPC: ***H04B 7/06*** (2006.01)

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Priority Data:

Title **(EN)** TRANSMISSION DIVERSITY COMMUNICATION DEVICE
(FR) DISPOSITIF DE COMMUNICATION POUR EMISSION EN DIVERSITE

Abstract: **(EN)** A plurality of antennas for transmission diversity at a base station are arranged into groups. Antennas are arranged such that the antennas in the same group transmit signals of strong fading correlation and that the fading correlation between groups is weak. Since the signals transmitted from the antennas in the same group have strong fading correlation, fading is not likely to change, and thus high control speed may not be necessary. On the other hand, the control between groups needs high speed. Therefore, mobile stations receiving signals from the base station transmit feedback information for controlling the fading between groups to the base station at high transmission speed, and feedback information in the same group is transmitted to the base station at low transmission speed.

(FR) Cette invention concerne une pluralité d'antennes pour émission en diversité au niveau d'une station de base, qui sont agencées en groupes, ceci de telle sorte que les antennes du même groupe transmettent des signaux à forte corrélation d'évanouissement, mais que la corrélation d'évanouissement entre les groupes soit faible. Comme le signal transmis par les antennes du même groupe ont une forte corrélation d'évanouissement, l'évanouissement a peu de chance de changer, faisant qu'il n'est pas nécessaire de disposer d'une vitesse de commande élevée. A l'inverse, la commande entre les groupes doit se faire à grande vitesse. Par voie de conséquence, les stations mobiles qui reçoivent des signaux de la station de base émettent des informations en retour pour la gestion de l'évanouissement entre les groupes et la station de base à grande vitesse, l'information en retour au sein du même groupe étant transmise à faible vitesse à la station de base.

Designated States: CN, JP, KR, US.
European Patent Office (EPO) (DE, FR, GB, IT).

Publication Language: Japanese (JA)

Filing Language: Japanese (JA)

(12)特許協力条約に基づいて公開された国際出願

(19) 世界知的所有権機関
国際事務局

(43) 国際公開日
2002年2月21日 (21.02.2002)

PCT

(10) 国際公開番号
WO 02/15433 A1

(51) 国際特許分類7: H04B 7/06, 7/26, H01Q 3/24

(21) 国際出願番号: PCT/JP00/05380

(22) 国際出願日: 2000年8月10日 (10.08.2000)

(25) 国際出願の言語: 日本語

(26) 国際公開の言語: 日本語

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(81) 指定国 (国内): CN, JP, KR, US.

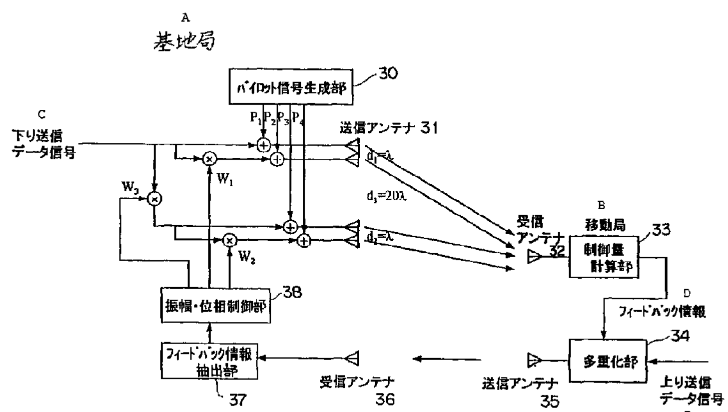
(84) 指定国 (広域): ヨーロッパ特許 (DE, FR, GB, IT).

添付公開書類:
— 国際調査報告書

２文字コード及び他の略語については、定期発行される各PCTガゼットの巻頭に掲載されている「コードと略語のガイダンスノート」を参照。

(54) Title: TRANSMISSION DIVERSITY COMMUNICATION DEVICE

(54) 発明の名称: 送信ダイバーシチ通信装置



A...BASE STATION
B...MOBILE STATION
C...DOWNSTREAM DATA SIGNAL
D...FEEDBACK INFORMATION
E...UPSTREAM DATA SIGNAL
30...PILOT SIGNAL GENERATOR
31...TRANSMISSION ANTENNA
32...RECEPTION ANTENNA
33...CONTROL QUANTITY CALCULATOR
34...MULTIPLEXER
35...TRANSMITTION ANTENNA
36...RECEPTION ANTENNA
37...FEEDBACK INFORMATION EXTRACTOR
38...AMPLITUDE/PHASE CONTROLLER

(57) **Abstract:** A plurality of antennas for transmission diversity at a base station are arranged into groups. Antennas are arranged such that the antennas in the same group transmit signals of strong fading correlation and that the fading correlation between groups is weak. Since the signals transmitted from the antennas in the same group have strong fading correlation, fading is not likely to change, and thus high control speed may not be necessary. On the other hand, the control between groups needs high speed. Therefore, mobile stations receiving signals from the base station transmit feedback information for controlling the fading between groups to the base station at high transmission speed, and feedback information in the same group is transmitted to the base station at low transmission speed.

[続葉有]

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(57) 要約:

送信ダイバーシチに使用する基地局の複数のアンテナをグループ分けする。グループ内のアンテナから送信される信号はフェージング相関が高くなるようにアンテナを配置する。グループ間のフェージング相関は低くなるように設定する。グループ内のアンテナから送信される各信号は、フェージング相関が高いので、フェージング変動を生じづらく、制御速度が遅くても十分である。一方、グループ間の制御は高速で行う必要がある。従って、基地局の信号を受信する移動局では、グループ間のフェージング変動を制御するためのフィードバック情報を高伝送速度で基地局にフィードバックし、グループ内のフィードバック情報は低伝送速度で基地局にフィードバックする。

明細書

送信ダイバーシチ通信装置

技術分野

本発明は、送信ダイバーシチ通信装置に関する。

背景技術

第３世代移動通信システムであるＷ－ＣＤＭＡにおける送信ダイバーシチでは、２本の送信アンテナを用いる方式が採用されている。

図１は、２本の送信アンテナを用いる場合の送信ダイバーシチシステムの構成例を示す図である。

２本の送信アンテナ１、２よりパイロット信号としてお互い直交するパイロットパターンP_1、P_2が送信され、移動局受信側では、それぞれの既知のパイロットパターンと受信パイロット信号との相関を取ることにより、基地局の各送信アンテナから移動局受信アンテナまでのチャネルインパルス応答ベクトル$\underline{h}_1$、$\underline{h}_2$を推定する。

これらチャネル推定値を用いて、以下の（１）式で示す電力Ｐを最大とする基地局各送信アンテナの振幅および位相制御ベクトル（重みベクトル）$\underline{w}=[w_1, w_2]$を制御量計算部１０において計算し、これを量子化する。そして、量子化された重みベクトルをフィードバック情報として多重化部１１において、上りチャネル信号に多重化して基地局側に伝送する。ただし、w_1、w_2の両方の値を伝送する必要はなく、$w_1=1$として求めた場合のw_2の値のみ伝送すれば良い。

$$P=\underline{w}^H H^H H\underline{w} \quad \cdots\cdots(1)$$

$H=[\underline{h_1},\underline{h_2}]$　　・・・・・(2)

ここで、$\underline{h}_1$、$\underline{h}_2$は、それぞれ送信アンテナ１および送信アンテナ２からのチャネルインパルス応答ベクトルであり、H^Hやw^Hの肩の添え字Hは、Hやwのエルミート共役をとることを示す。インパルス応答の長さをＬとすると、次式で表される。

$\underline{h_i}=[h_{i1},h_{i2},\cdots,h_{iL}]^T$・・・・・・(3)

従って、上記式（１）の計算は、送信アンテナが２の場合、以下のような代数計算に基づいて行われる。

$$H=\begin{bmatrix} h_{11} & h_{21} \\ h_{12} & h_{22} \\ \vdots & \vdots \end{bmatrix}\quad、\quad \underline{w}=[w_1,w_2]^T \quad、よって、\ Hw=\begin{bmatrix} h_{11}w_1+h_{21}w_2 \\ h_{12}w_1+h_{21}w_2 \\ \vdots \end{bmatrix}$$

ハンドオーバ時には（１）式の代わりに次式を最大とする重みベクトル$\underline{w}$を計算する。

$P=\underline{w^H}(H_1^H H_1+H_2^H H_2+\cdots)\underline{w}$　　・・・・・・(4)

ここで、H_kはｋ番目の基地局からの信号のチャネルインパルス応答である。

そして、移動局から送信されてくるw_2を（ここで、$w_1=1$となっているとする）送信側のフィードバック情報抽出部１２で受信信号の中から抽出し、振幅・位相制御部１３において、w_2を送信アンテナ２から送出されるべきデータ信号に乗算する。これにより、受信側で受け取る送信アンテナ１と送信アンテナ２からの信号の振幅及び位相の劣化を送信側で予め修正してから、送出することが出来る。

Ｗ－ＣＤＭＡでは、重み係数w_2を１ビットに量子化するモード１と、４ビットに量子化するモード２の２通りの方法が規定されている。モード１では１ビットのフィードバック情報を毎スロット伝送して制御するため、制御速度が速い反面、量子化が粗いため正確な制御が出来ない。一方、モード２では４ビ

ットの情報で制御するため、より精度の高い制御が出来る反面、各スロットで１ビットずつ伝送して４スロットで１ワードのフィードバック情報を伝送するため、フェージング周波数が高い場合にはこれに追従出来ずに、振幅・位相制御特性が劣化する。このように、フィードバック情報を伝送する、移動局から送信局への上りチャネルの信号伝送レートが限られている場合、制御精度とフェージングに対する追従速度はトレードオフの関係にある。

Ｗ－ＣＤＭＡの Release-99 規格では、フィードバック情報伝送による上りチャネル伝送効率の低下を回避するため、送信アンテナ数として２本より多い場合は考慮されていない。しかしながら、フィードバック情報の増加や更新速度の低減を許容すれば、３本以上への拡張も可能である。特に、現在では、送信アンテナ数を４本とする場合が盛んに研究開発されている。

セルラ移動通信システムの無線基地局に閉ループ送信ダイバーシチ方式を適用すると、各送信アンテナからの信号が独立のフェージングを受けた後、理想的には移動局アンテナ位置に於いて同相合成されるため、送信アンテナ数に応じたダイバーシチ利得が得られることに加えて、合成による利得向上が得られる。このため、受信特性が向上すると共に、１つのセルに収容できるユーザ数を増大することができる。ここでいう理想的とは、フィードバック情報の伝送誤り、制御遅延、チャネル応答推定誤差、制御量の量子化誤差がない場合を言う。実際にはこれらの要因により理想的な場合に比べて特性は劣化する。

送信アンテナ数に応じたダイバーシチ利得を得るためには、フェージング相関が十分小さくなるようにアンテナ間隔を大きく取る必要がある。一般的にセルラ移動通信システムの無線基地局でフェージング相関を十分小さく抑えるためにはアンテナ間隔を２０波長程度とる必要がある。２ＧＨｚ帯では１波長は約１５ｃｍであるから約３ｍ間隔でアンテナを設置することになる。このため、送信アンテナ数を増加させるとアンテナ設置に必要な面積が大きくなり、建物

の屋上等に設置する場合に設置が困難となる問題がある。また、ダイバーシチ利得は送信アンテナ数の増加と共に飽和していくため、あまり送信アンテナ数を増やしても大きなダイバーシチ利得の改善は得られない。

また、送信アンテナ数を増加させると、各アンテナについてフィードバック情報を伝送しなくてはならないため、フィードバックすべき情報量が増え、フィードバック情報伝送のために上りチャネルの伝送効率が低下したり、送信ダイバーシチの制御が高速なフェージングに追従できなくなって特性劣化を引き起こすといった問題がある。

発明の開示

本発明の課題は、送信アンテナ数を増加させた場合に、上りフィードバック情報の増加を抑えられ、フェージング周波数の高い場合における特性劣化が少なく、基地局アンテナ設置スペースが小さい送信ダイバーシチ通信装置を提供することである。

本発明の送信ダイバーシチ通信装置は、移動局からの情報を基に送信信号の制御を行う、送信ダイバーシチ方式を適用した基地局を有する送信ダイバーシチ通信装置であって、複数本ずつのアンテナが複数のグループに分けられ、該グループ内のアンテナは該グループ内のアンテナ間のフェージング相関が高くなるように近接して、各グループのアンテナ群は該グループ間のフェージング相関が低くなるよう互いの距離が大きくなるように設置されたアンテナ手段と、移動局から送信されてくる該グループ内アンテナの制御に関する送信速度の遅い第１の制御情報と、該各グループのアンテナ群間の制御に関する送信速度の速い第２の制御情報とを受け取って、該アンテナ手段の送信する信号の位相を制御する制御手段とを備えることを特徴とする。

本発明によれば、アンテナ本数が増えることにより、従来の２本の時と同様

の方法で、送信信号の制御を閉ループ送信ダイバーシチ方式に適用しようとすると、移動局から基地局に送信すべき情報量が増えてしまうため、フェージング変動の追従性が悪くなり、送信信号の制御性能が落ちてしまうことを防ぐことが出来る。

特に、本発明では、基地局のアンテナをグループ化し、グループ内ではフェージング相関が高く、各グループ間ではフェージング相関が低くなるように設定するので、高速で制御情報を移動局から基地局に送る必要があるのは、各グループ間の送信信号の制御についてのみであり、グループ内のアンテナからの送信信号制御は比較的ゆっくりでよくなる。従って、限られた移動局から基地局への上り回線の伝送速度を有効に利用して、送信ダイバーシチ方式の性能を向上することが出来る。

図面の簡単な説明

図１は、２本の送信アンテナを用いる場合の送信ダイバーシチシステムの構成例を示す図である。

図２は、本発明の原理を説明するシステム構成図である。

図３は、本発明の実施形態に従う基地局の送信アンテナの構成例を示す図である。

図４は、本発明の一実施形態の構成を示す図である。

図５は、本実施形態における下りパイロット信号パターンの例を示す図である。

図６は、本実施形態に従う基地局送信アンテナの構成例及びアンテナの制御情報について説明する図である。

図７は、マクロセル環境における基地局で観測される到来信号の角度分散Δφは約３度程度ときの包絡線相関係数の様子を示す図である。

図８は、本実施形態におけるフィードバック情報の伝送フォーマットの例を示す図（その１）である。

図９は、本実施形態におけるフィードバック情報の伝送フォーマットの例を示す図（その２）である。

図１０は、本実施形態におけるフィードバック情報の伝送フォーマットの例を示す図（その３）である。

図１１は、本実施形態におけるフィードバック情報の伝送フォーマットの例を示す図（その４）である。

図１２は、図８～図１１のフォーマットに従って基地局にフィードバック情報を送信する移動局の構成例を示した図である。

図１３は、本発明の第２の実施形態における基地局の構成例を示す図である。

図１４は、第２の実施形態におけるグループ内アンテナの位相差制御方法を説明する図である。

図１５は、本発明の第３の実施形態を説明する図である。

発明を実施するための最良の形態

本発明は、セルラ移動通信システム無線基地局に複数のアンテナ素子を設け、同一の送信データ信号に移動局からのフィードバック情報に基づいて異なる振幅及び位相制御を行った後、それぞれ異なるアンテナを用いて送信を行い、移動局側では該振幅及び位相制御量を下りパイロット信号を用いて決定し、該振幅及び位相制御量を表すフィードバック情報を上りチャネル信号に多重して基地局側に伝送する閉ループ送信ダイバーシチ方式に関する技術である。

図２は、本発明の原理を説明するシステム構成図である。

基地局のパイロット信号生成部２０でＮ個のお互い直交するパイロット信号$P_1(t)$、$P_2(t)$、・・・、$P_N(t)$を生成し、それぞれ異なる送信アンテ

ナ２１を用いて送信する。Ｎは送信アンテナ数である。これらパイロット信号間には以下の関係がある。

$$\int P_i(t)P_j(t)dt = 0 \qquad (i \neq j)$$

各パイロット信号は、それぞれフェージングによる振幅及び位相変動を受け、これらの合成信号が移動局受信アンテナ２２に入力される。移動局受信機では受信パイロット信号に対して$P_1(t)$、$P_2(t)$、・・・、$P_N(t)$との相関をそれぞれ求めることにより、各パイロット信号のチャネルインパルス応答ベクトル$\underline{h}_1$、$\underline{h}_2$、・・・、$\underline{h}_N$を推定する。

これらチャネルインパルス応答ベクトルを用いて、以下の（５）式（上記（１）式に同じ）で示す電力Ｐを最大とする基地局各送信アンテナの振幅および位相制御ベクトル（重みベクトル）$\underline{w} = [w_1, w_2, \cdots, w_N]^T$を制御量計算部２３で計算し、これを量子化してフィードバック情報として多重化部２４より、上りチャネル信号に多重化して基地局側に伝送する。ただし、この場合でも$w_1 = 1$として求めた場合のw_2、w_3、・・・、w_Nの値を伝送すればよい。

$$P = \underline{w}^H H^H H \underline{w} \qquad \cdots\cdots(5)$$

$$H = [\underline{h}_1, \underline{h}_2, \cdots, \underline{h}_N] \qquad \cdots\cdots(6)$$

ここで、$\underline{h}_i$は送信アンテナｉからのチャネルインパルス応答ベクトルである。インパルス応答の長さをＬとすると、次式で表される。

$$\underline{h}_i = [h_{i1}, h_{i2}, \cdots, h_{iL}]^T \qquad \cdots\cdots(7)$$

ハンドオーバ時には（５）式の代わりに次式を最大とする重みベクトル$\underline{w}$を計算する。

$$P = \underline{w}^H (H_1^H H_1 + H_2^H H_2 + \cdots)\underline{w} \qquad \cdots\cdots\cdot(8)$$

ここで、H_kはｋ番目の基地局からの信号のチャネルインパルス応答であり、上記（４）式と同じである。

このようにして得られた重みベクトルは、移動局の多重化部２４から上り送

信データ信号に多重されて基地局の受信アンテナに送信される。基地局では、受信アンテナで受信したフィードバック情報をフィードバック情報抽出部２５において、抽出し、該フィードバック情報に含まれる重みベクトルを用いて、各送信アンテナ２１から送信される信号の振幅及び位相を振幅・位相制御部２６において制御する。このようにして、振幅及び位相が制御された後の信号が基地局の送信アンテナ２１から送出されると、ちょうどフェージングによる振幅及び位相の変動が補償された形で移動局によって受信されるので、最適な受信を行うことが出来る。また、フェージングは時間が経過する毎に変化するので、フェードバック情報の生成と送信はリアルタイムで行う必要がある。しかし、移動局から基地局への上り送信データ信号の送信フォーマットや送信速度は予め決められているので、多くの情報を送ろうとすると時間がかかってしまうために、フェージングの変化を追従できなくなってしまう。また、フェージングの変化に追従できるようにするためには、フィードバック情報の送信速度を速くしてやる必要があるが､上り回線の通信速度が決められていることから、高速で送信ダイバーシチの制御をするために短い周期で新しい情報を次々と送ろうとすると、１回の送信に含まれる情報量が少なくなってしまい（量子化が粗くなってしまい)、精度の良い制御が出来なくなってしまう。

本発明の実施形態では、重みベクトルを構成する各係数値を、それぞれのアンテナから送られてくる信号に対して同じ周期で計算してフィードバックするのではなく、それぞれ異なる周期で計算およびフィードバックを行うようにする。

以下に具体的に説明する。

図３は、本発明の実施形態に従う基地局の送信アンテナの構成例を示す図である。

基地局において、図３に示すように送信アンテナを、複数のアンテナで構成

される複数のグループに分け、各グループ内の送信アンテナはフェージング相関が高くなるように近接して配置し、グループ間はフェージング相関が低くなるように互いの距離を離して設置する。フェージング相関とは、異なるアンテナから送信された信号が受信側で受信した際に、どのくらい同じようなフェージングを受けているかを示す数値である。フェージングは、建物による反射や移動体に反射することによるドップラー効果などによって引き起こされるが、送信信号を送信する送信アンテナがそれぞれ近接していれば、それぞれから送信された信号は、同じような経路を辿って移動局に受信されると考えられ、従って、同じようなフェージングを受けることになる。このような場合、フェージング相関が高いという。一方、信号を送信するアンテナが互いに離れていると、それぞれから送信される信号は、移動局によって受信されるまでに、異なる経路を辿っていると考えられ、従って、異なるフェージングを受けて移動局によって受信されることになる。このような場合を、フェージング相関が低いという。

移動局において、グループ間のアンテナ制御量を、グループ内のアンテナ制御量より速い周期で計算し、これらをフィードバック情報として基地局側に伝送する。同一グループ内の基地局送信アンテナからの信号はフェージング相関が高いため、ほぼ同一のフェージングを受けるが、移動局の基地局に対する角度に応じた位相差をもって移動局の受信アンテナに到達する。従って、同一グループ内の基地局送信アンテナからの信号を用いて推定したチャネル応答推定値は、この移動局の基地局に対する角度に応じた位相差を有する。移動局の移動に伴ってこれらの値は変化するが、フェージング変動に比べてゆっくりと変動することが特徴とである。各グループ内で１つのアンテナを基準アンテナとし、基準アンテナ以外の当該グループ内アンテナの制御量を、この基準アンテナの制御量で正規化（基準アンテナの制御量を基準として、これからの相対値

を使用）する。この正規化されたグループ内アンテナ制御量は移動局の移動に伴いゆっくりと変動する。従って、制御周期を比較的長くすることが可能である。

一方、異なるグループの基地局送信アンテナからの信号間のフェージング相関は低いため、それぞれ独立なフェージングを受けて移動局の受信アンテナに到達する。従って、異なるグループの基準アンテナからの信号を用いて推定したチャネル応答推定値（チャネルインパルス応答ベクトル）は、それぞれ独立なフェージング変動によって高速に変動することが特徴である。ある特定の１つのグループの基準アンテナ制御量で他のグループの基準アンテナ制御量を正規化したものを、グループ間アンテナ制御量と定義する。このグループ間アンテナ制御量はそれぞれ独立なフェージング変動によって高速に変動するため、正確な制御を行うためには短い周期で制御を行う必要がある。

移動局は、どの信号がどのグループのどのアンテナから来たものであるかを認識する必要があるが、これは、予めアンテナとそのアンテナから送信されるパイロット信号とを対応付けておけば十分である。各パイロット信号は互いに直交しているので、受信側では、その信号がどのアンテナから送信されたものかをパイロット信号を調べることによって正確に認識することができる。

図３に示したグループ間アンテナの制御量$F_{1、m}$、及びグループ内アンテナ制御量$G_{m、k}$は次式のように求める。ここで、Ｎは総アンテナ数、Ｍはアンテナグループ数、Ｋ＝Ｎ／Ｍは各グループのアンテナ数である。＊は複素共役を表す。

全体基準アンテナ：アンテナ＃１

グループ内基準アンテナ：アンテナ＃（(ｍ－１）Ｋ＋１）（ｍ＝１、・・・、M)

$$F_{1,m}=\frac{w_{(m-1)K+1}}{w_1} \quad (m=1、・・・、M) \quad ・・・・・(9)$$

$$G_{m,k}=\frac{w_{(m-1)K+k+1}}{w_{(m-1)K+1}} \quad (M=1、・・・、M、k=1、・・・、K)・・・(10)$$

グループ内ではフェージング相関が高いため、｜$G_{m、k}$｜＝１とおくことができる。すなわち、グループ内では、フェージングによる振幅の変化は小さく、位相の変化のみを考えれば十分であると考えることが出来る。また、総送信電力を一定（＝１．０）に保つために次のように$F_{1、m}$の正規化を行う。

$$F'_{1,m}=\frac{F_{1,m}}{\sqrt{\frac{1}{KM}\sum_{j=1}^{M}\left|F_{1,j}\right|^2}}$$

次に、フェージングの変動速度について述べる。

フェージング変動速度はドップラー周波数で表される。

$$f_d=\frac{\nu}{\lambda}$$

νは移動局の移動速度、λはキャリアの波長である。例えば、キャリア周波数２ＧＨｚで移動局の移動速度が６０ｋｍ／ｈの場合、f_dは約１１１Ｈｚになる。一方、到来波の到来角度は移動局の移動に合わせて変化するが、例えば２００ｍ先を速度２００ｋｍ／ｈで移動した場合の到来角度の変化は約１５度／秒である。このようにフェージング変動速度は到来角度変動速度に対して２桁から３桁程度高速である。Ｗ－ＣＤＭＡの規格では、スロット長は６６６．７μｓであり、フィードバック情報の更新速度は１５００Ｈｚとなる。従って、フェージング変動に対しては、スロット毎に更新しないと追従特性が劣化するが、到来角度変動に対してはスロット毎に更新する必要な無く、例えば、１５スロット（＝１フレーム）毎に更新を行っても問題がない。

上述した制御情報の変動速度の違いを利用することにより、性能を劣化させることなくフィードバック情報量を削減することが出来る。すなわち、高速に変動するグループ間アンテナ制御量は短い周期で更新、フィードバックを行い、これと比較して変動速度が遅い各グループ内アンテナ制御量はこれより長い周期で更新、フィードバックを行う。すなわち、フェージング変動が生じると考えられる、フェージング相関の低いグループ間のダイバーシチ制御は、フィードバック情報のデータ速度と比べても変化が速いので、更新頻度を速くするが、フェージング変動が生じず、単に到来角度の変動による変化のみが生じると考えられる、フェージング相関の高いグループ内のダイバーシチ制御は、フェードバック情報のデータ速度と比べてもゆっくりなので、更新頻度を遅くする。

また、各グループ内アンテナ制御量は移動局の基地局に対する角度に応じた値を持つため、セル半径がある程度大きいマクロセルシステムでは、到来角度のずれはほとんど無視できるほど小さくなる。このため、ある特定のグループ内アンテナ制御量を他のグループのグループ内アンテナ制御量として用いることも出来る。すなわち、ある１つのグループのグループ内制御情報のみを伝送し、これを用いてその他のグループ内のアンテナの制御も行うことにより、フィードバック情報量を更に削減することが出来る。

図４は、本発明の一実施形態の構成を示す図である。

アンテナ数N＝４、アンテナグループ数M＝２の場合について説明する。N＝４個のパイロット信号P_1（t）、P_2（t）、P_3（t）、P_4（t）をパイロット信号生成部３０において生成し、各送信アンテナより送信する。これらパイロット信号には互いに直交するビットシーケンスを用いる。

パイロット信号は、送信アンテナ３１から移動局に向けて送出される。移動局では、４つの送信アンテナ３１から送信された４つのパイロット信号を受信アンテナ３２で受信し、制御量計算部３３において、パイロット信号を用いて、

各送信アンテナ３１から送信された信号のチャネル推定を行う。この結果、各アンテナからの信号にたいしてチャネルインパルス応答ベクトルが得られ、（５）式を最大にするような重みベクトルを算出する。この重みベクトルを算出する方法は、既に公知であるので説明を省略する。制御量計算部３３は、重みベクトルが算出できると、これをフィードバック情報として、多重化部３４に転送する。多重化部３４では、上り送信データ信号にフィードバック情報を多重して、送信アンテナ３５から送出する。基地局では、受信アンテナ３６で移動局からの信号を受信し、フィードバック情報抽出部３７でフィードバック情報を抽出する。抽出されたフィードバック情報は、振幅・位相制御部３８に入力され、フィードバック情報に含まれる重み係数w_1、w_2、w_3が対応するアンテナの下り送信データ信号に乗算されて、送信アンテナ３１からこの下り送信データ信号が送出される。このように、本実施形態においては、基地局と移動局を含めて、送信ダイバーシチ制御を行うための閉ループが形成されている。

図５は、本実施形態における下りパイロット信号パターンの例を示す図である。

図５のパイロット信号P_1～P_4は、互いに対応する符号を乗算し、これをパイロット信号パターン全体について加算すると、結果として“０”が得られるようになっている。すなわち、パイロット信号P_1～P_4は、互いに直交する符号列となっている。

各パイロット信号はそれぞれフェージングによる振幅及び位相変動を受け、これらの合成信号が移動局受信アンテナに入力される。移動局受信機では受信パイロット信号に対して、予め移動局側で保持しているパイロット信号パターンP_1（t）、P_2（t）、P_3（t）、P_4（t）との相関をとって平均することにより、各パイロット信号のチャネル応答推定値$\underline{h}_1$、$\underline{h}_2$、$\underline{h}_3$、$\underline{h}_4$を求める

ことが出来る。

図６は、本実施形態に従う基地局送信アンテナの構成例及びアンテナの制御情報について説明する図である。

図６（ａ）は、基地局送信アンテナ構成を示す。アンテナＡＮＴ１とアンテナＡＮＴ２をグループ１、アンテナＡＮＴ３とアンテナＡＮＴ４をグループ２とする。また、アンテナＡＮＴ１とアンテナＡＮＴ３を各グループにおける基準アンテナとする。更に、アンテナＡＮＴ１を全グループの基準アンテナとする。アンテナＡＮＴ１とアンテナＡＮＴ２、及びアンテナＡＮＴ３とアンテナＡＮＴ４はそれぞれ１波長分の長さ離れて配置されている。アンテナＡＮＴ１とアンテナＡＮＴ３、及びアンテナＡＮＴ２とアンテナＡＮＴ４はそれぞれ２０波長分の長さ離れて配置されている。

ここで、基地局アンテナの空間相関特性について説明する。

移動局からの信号の到来角度が分散Δφで一様分布している場合、到来波の包絡線相関係数は以下の式で表される。ここでｄはアンテナ間隔である。

$$\rho = \left(\frac{\sin X}{X}\right)$$

$$X = \frac{\pi d \Delta\phi}{\lambda}$$

マクロセル環境における基地局で観測される到来信号の角度分散Δφは約３度程度であるので、このときの包絡線相関係数は、図７のようになる。この図より、ｄ≃ １９λで無相関となることが分かる。従って、本発明においてアンテナグループ間の距離を１９波長以上程度とすることにより、フェージング相関を低くすることができる。また、各グループ内のアンテナ間距離を１波長以下とすることにより、フェージング相関を高くすることが出来る。

ただし、フェージング相関は、アンテナの配置されている高さやアンテナの大きさなど様々なファクターによって影響を受ける。従って、グループ内アン

テナの配置間隔は、ほぼ到来波の波長程度でよいが、グループ間アンテナの間隔は、各状況に合わせてフェージング相関がほぼ“0”になるように当業者によって設定されるべきものである。

図６に戻って説明する。なお、以下の説明では振幅の制御は行わず、位相のみの制御を行う方法について説明する。すなわち、$w_i = a_i e^{j\phi_i}$と表した場合に$a_i = 1$として位相量ϕ_iのみの制御を行う。図６（ｂ）に示すように、アンテナＡＮＴ１を基準としたアンテナＡＮＴ２の制御量ϕ_1、アンテナＡＮＴ３を基準としたアンテナＡＮＴ４の制御量ϕ_2、及びアンテナＡＮＴ１を基準としたアンテナＡＮＴ３の制御量ϕ_3をそれぞれ量子化し、これらをフィードバック情報として基地局に伝送する。それぞれ１ビットで量子化を行う場合、例えば、以下のようにする。

$$\begin{aligned} &\frac{-\pi}{2} < \phi_i \le \frac{\pi}{2} \Rightarrow \phi_i^Q = 0 \\ &\frac{\pi}{2} < \phi_i \le \frac{3\pi}{2} \Rightarrow \phi_i^Q = \pi \end{aligned} \qquad \cdots\cdots\cdot (11)$$

ここで、ϕ_i^Qは、量子化された制御量である。

図８～図１１は、本実施形態におけるフィードバック情報の伝送フォーマットの例を示す図である。

$\phi_i^Q = 0$の場合は、フィードバック情報は$b_i = 0$、$\phi_i^Q = \pi$の場合はフィードバック情報は$b_i = 1$とする。このフィードバック情報は図８に示すようにb_3の伝送速度をb_1及びb_2の伝送速度より高くなるように上りチャネルに多重化して基地局に伝送する。ここでは、Ｗ－ＣＤＭＡのフレームフォーマットにのっとり、１５のスロットで構成された１０ｍｓ長の１フレームを示している。本伝送フォーマットでは、各スロットで１ビットのフィードバック情報を伝送する。フォーマット１はb_1及びb_2を１フレームに各１回伝送する場合、フォーマット２はb_1及びb_2を１フレームに各２回伝送する場合の例である。

基地局では上りチャネルで受信されたフィードバック情報を用いて各送信アンテナの位相制御を行う。直前の受信スロットで受信したフィードバック情報で対応するアンテナを直接制御する。このとき、それ以外のアンテナは時間的に最近のフィードバック情報を保持して制御に用いる。

ただし、図６（ａ）のＡＮＴ４については、ｄ２によってのみ制御がなされるのではなく、ＡＮＴ３の制御量ｄ３によっても制御される。すなわち、ｄ３で頻繁に制御されると共に、ｄ２によって遅く制御されることになる。これは図６（ｂ）のＡＮＴ４についても同様である。

フィードバック情報のフィルタリングを行い、伝送誤りや量子化誤差の低減を図ることも出来る。例えば、フィルタリングの例としては、直前の受信スロットで受信したフィードバック情報による制御量とそれ以前の時間的に最近の当該フィードバック情報による制御量の２つを平均した値を、実際に用いる制御量とする方法が考えられる。

グループ内アンテナ制御量のフィードバック情報は、送る度に更新された制御量を送るが、別の方法として、例えば同一フレーム内では同じフィードバック情報を繰り返し送ることも可能である。この場合、基地局ではフレーム内で受信された当該フィードバック情報を合成することにより、伝送誤りを低減することが出来る。

また、各グループ内アンテナ制御量は移動局の基地局に対する角度に応じた値を持つため、セル半径がある程度大きいマクロセルシステムではグループ内の到来角度のずれはほとんど無視できるほど小さくなる。このため、各グループで同一のグループ内アンテナ制御量を用いて制御を行っても問題が無い。このことを利用して、ある１つのグループのグループ内制御情報のみを伝送し、これを用いてその他のグループの制御も行うことにより、フィードバック情報量を更に削減することが出来る。

図９にグループ内制御情報としてb_1のみを伝送する場合のフィードバック情報伝送フォーマットを示す。フォーマット３はb_1を１フレームに２回伝送する場合、フォーマット４はb_1を１フレームに４回伝送する場合の例である。

本実施例においても、グループ内アンテナ制御量のフィードバック情報は、送る度に更新された制御量を送っても良いし、別の方法として、例えば同一フレーム内では同じフィードバック情報を繰り返し送ることも可能である。

別の伝送フォーマットとして、移動局で求めた制御量を複数ビットで量子化する場合を示す。

図１０は、b_1を３ビット、b_3を４ビットで量子化して伝送する場合のフィードバック情報伝送フォーマットを示す図である。また、図１１の表１、２にグループ間アンテナ制御量のフィードバック情報b_3との対応例、表３にグループ内アンテナ制御量のフィードバック情報b_1との制御量の対応例を示す。

ここでは、図９のフォーマットを適用して、グループ内アンテナ制御量のフィードバック情報としてのb_1のみを伝送している。図１１の表１、２から明らかなように、フィードバックビットb_3は、位相の制御量を示すb_3（３）～b_3（１）の３ビットと、振幅の制御量を示すb_3（０）の１ビットの計４ビットからなっており、図１０のフォーマット５によれば、１フレームに３ワード含まれている。一方、フィードバックビットb_1は、b_1（２）～b_1（０）までの３ビットで、位相の制御量を示している。そして、図１０のフォーマット５によれば、フィードバックビットb_1は、１フレームに、その３ビットが分散されて配置されており、全体で１ワードを形成するように構成されている。

図１２は、図８～図１１のフォーマットに従って基地局にフィードバック情報を送信する移動局の構成例を示した図である。

移動局は、基地局からの信号を受信アンテナ４０で受信すると、受信信号を２つに分岐し、一方をデータチャネル逆拡散部４１に、他方をパイロットチャ

ネル逆拡散部４４に入力する。データチャネル逆拡散部４１では、データチャネルの信号を逆拡散し、チャネル推定部４２と受信機４３に入力する。受信機４３は、チャネル推定部４２のチャネル推定結果に基づいて下り受信データ信号を復元し、ユーザに音声あるいはデータとして提示する。一方、パイロットチャネル逆拡散部４４は、受信信号をパイロットチャネルの拡散符号で逆拡散し、チャネル推定部４５に入力する。チャネル推定部４５において、逆拡散された信号に対し、各パイロット信号パターンとの相関がとられ、各送信アンテナから移動局までのチャネル推定値$H=[\underline{h}_1、\underline{h}_2、\underline{h}_3、\underline{h}_4]$が得られる。制御量計算部４６では、このチャネル推定値より重みベクトルを求め、送信すべきフィードバック情報を決定する。このフィードバック情報は、多重化部４７において、上り制御チャネルに多重化され、データ変調部４８において、変調され、拡散変調部４９において、拡散変調された後、送信アンテナ５０から基地局に伝送される。

図１３は､本発明の第２の実施形態における基地局の構成例を示す図である。また、図１４は、第２の実施形態におけるグループ内アンテナの位相差制御方法を説明する図である。

なお、図１３において、図４と同じ構成要素には同じ参照符号を付し、説明を省略する。

本実施形態においては、基地局において、グループ内アンテナ制御情報として、上りフィードバック情報及び上りチャネルの到来方法推定結果の双方を用いる。基地局では、アレーアンテナ（送信ダイバーシチで使用する複数のアンテナの列：送受信アンテナ６０）で受信された上り受信信号から、到来方向推定部６２、６３で受信信号の到来方向を推定する。到来方向は移動局の基地局に対する角度に強く依存するため、下りの送信ビームの方向（アンテナから送信する電波の強度の大きい方向）を、この上り受信信号到来方向に設定する方

法が知られている。ただし、上りと下りの周波数が異なるシステムにおいては伝播環境により必ずしもこの仮定が成り立たない場合もある。

上りフィードバック情報は、送受信アンテナ６０において受信された後、受信処理部６１において逆拡散などの処理を受け、フィードバック情報抽出部３７に送られる。フィードバック情報抽出部３７において抽出された制御量は、振幅・位相制御部３８’において、到来方向推定値と比較され、上り回線から受信した制御量を使用するか、到来方向推定値を使用するかが決定され、送信信号の振幅・位相制御が行われる。

本実施形態では、図１４に示すようにグループ内位相差情報の上りフィードバック情報による制御量が、上りチャネルの到来方向推定結果θと関連があることを利用して、フィードバック情報による制御量が上りチャネルの到来方向推定結果θを中心とした一定の範囲［$\theta-\Delta$、$\theta+\Delta$］にない場合に、到来方向推定結果θのみを用いて制御を行う。すなわち、フィードバック情報による制御量が到来方向推定結果θからあまり離れた値を示している場合には、移動局から基地局にフィードバック情報が送信されてくる間に、ビット誤りなどを生じ、不正確な情報となったと判断し、フィードバック情報を破棄し、到来方向推定結果θを用いて、位相の制御のみを行うようにしたものである。

また、別の方法として、グループ内位相差情報の上りフィードバック情報による制御量を所定時間サンプルし、この制御量のサンプルの分散が大きいと判断された場合には（例えば、すなわち、予め定められた所定の閾値よりも分散が大きい場合には)、フィードバック情報を使用せずに到来方法推定結果θのみを用いて制御を行う。

図１５は、本発明の第３の実施形態を説明する図である。

なお、図１５において、図４と同じ構成には、同じ参照符号を付し、説明を省略する。

パイロット信号P_2及びP_4の送信電力をP_1及びP_3の送信電力より小さく設定する。本実施形態ではパイロット信号P_2及びP_4に係数α（$0<\alpha\leqq1$）を乗じることにより実現している。パイロット信号P_2及びP_4はチャネルインパルス応答ベクトル$\underline{h}_2$、$\underline{h}_4$を推定するために必要であるが、$\underline{h}_2$、$\underline{h}_4$はそれぞれ$\underline{h}_1$、$\underline{h}_3$と高いフェージング相関があるため、これらで正規化した$\underline{h}_2/\underline{h}_1$及び$\underline{h}_4/\underline{h}_3$は移動局の基地局に対する角度に依存した値となる。これらの値はフェージング変動に比べて時間的にゆっくり変動するため、移動局側において受信電力が低くなっても、パイロット信号P_2及びP_4の長時間平均を取ることにより推定精度を向上させることができる。ϕ_1及びϕ_2は以下のように求める。

$$\phi_1=\underline{h}_2/\underline{h}_1,\phi_2=\underline{h}_4/\underline{h}_3 \quad \cdots\cdots(12)$$

パイロット信号P_2及びP_4の送信電力を低く設定することにより、これらパイロット信号によるデータ信号への干渉を低く抑えることができるため、伝送容量を増大させることが出来る。

$\underline{h}_2/\underline{h}_1$及び$\underline{h}_4/\underline{h}_3$は移動局の基地局に対する角度に依存した値であり、フェージング変動に比べて時間的にゆっくり変動するため、受信電力が低くなっても長時間平均を取ることにより推定精度を向上させることが出来る。例えば、第nスロットにおける推定値ϕ_1（n）、ϕ_2（n）及びϕ_3（n）は以下のように求める。ここで、Nは、ϕ_1（n）、ϕ_2（n）の推定における平均スロット数である。

$$\phi_1(n)=\frac{1}{N}\sum_{i=0}^{N-1}\frac{\underline{h}_2(n-i)}{\underline{h}_1(n-i)}$$

$$\phi_2(n)=\frac{1}{N}\sum_{i=0}^{N-1}\frac{\underline{h}_4(n-i)}{\underline{h}_3(n-i)}$$

$$\phi_3(n)=\frac{\underline{h}_3(n)}{\underline{h}_1(n)}$$

このように、ϕ_1及びϕ_2をϕ_3のN倍の時間（スロット数）平均をとって求める場合、$\alpha=1/N$としてもϕ_3と同等の推定精度を得ることが出来る。すなわち、N＝4とした場合、$\alpha=1/4$程度とすることが出来る。

産業上の利用可能性

制御情報の変動速度の違いを利用することにより、送信アンテナ数を増加させた場合に、

・ 上りフィードバック情報の増加が抑えられる。

・ フェージング周波数が高い場合における特性劣化が少ない。

・ 基地局アンテナ設置スペースを小さくできる。

といった効果が得られる。

請求の範囲

１．移動局からの情報を基に送信信号の制御を行う、送信ダイバーシチ方式を適用した基地局を有する送信ダイバーシチ通信装置であって、

複数本ずつのアンテナが複数のグループに分けられ、該グループ内のアンテナは該グループ内のアンテナ間のフェージング相関が高くなるように近接して、各グループのアンテナ群は該グループ間のフェージング相関が低くなるよう互いの距離が大きくなるように設置されたアンテナ手段と、

移動局から送信されてくる該グループ内アンテナの制御に関する送信速度の遅い第１の制御情報と、該各グループのアンテナ群間の制御に関する送信速度の速い第２の制御情報とを受け取って、該アンテナ手段の送信する信号の位相を制御する制御手段と、

を備えることを特徴とする送信ダイバーシチ通信装置。

２．前記移動局は、前記基地局から送られてくるパイロット信号を用いて、前記位相の制御量を決定することを特徴とする請求の範囲第１項に記載の送信ダイバーシチ通信装置。

３．前記制御手段は、前記位相の制御に加えて、振幅の制御も行うことを特徴とする請求の範囲第１項に記載の送信ダイバーシチ通信装置。

４．前記移動局は、前記基地局から送られてくるパイロット信号を用いて前記位相及び振幅の制御量を決定することを特徴とする請求の範囲第３項に記載の送信ダイバーシチ通信装置。

５．前記移動局は、基地局から受信したパイロット信号と、移動局側で既知のパイロット信号との相関を取ることによって、各アンテナから移動局までのチャネル応答を推定し、このチャネル応答推定値を用いて、前記制御量を求めることを特徴とする請求の範囲第２項または第４項に記載の送信ダイバーシチ通信装置。

６．前記移動局は、前記アンテナ手段の各グループ内のアンテナについては、該グループの基準アンテナからのチャネル応答推定値に対する差分情報、及び前記各グループのアンテナ群については、特定のグループの基準アンテナに対する各グループの基準アンテナのチャネル応答推定値の差分情報を、それぞれ前記第１の制御情報及び第２の制御情報として前記基地局に送信することを特徴とする請求の範囲第１項に記載の送信ダイバーシチ通信装置。

７．前記移動局は、前記各グループのアンテナ群についての制御情報と、ある一つのグループのグループ内のアンテナに対する制御情報とを、それぞれ前記第２及び第１の制御情報として前記基地局に送信することを特徴とする請求の範囲第１項に記載の送信ダイバーシチ通信装置。

８．前記制御手段は、前記第１の制御情報及び第２の制御情報に加え、上りチャネルの信号の到来方向推定結果を用いて、前記基地局からの信号の送信制御を行うことを特徴とする請求の範囲第１項に記載の送信ダイバーシチ通信装置。

９．前記第１の制御情報および第２の制御情報から得られる送信信号の制御量が、前記上りチャネルの信号の到来方向推定結果を中心とした一定範囲内にない場合には、該到来方向推定結果を用いて送信制御を行うことを特徴とする請

求の範囲第８項記載の送信ダイバーシチ通信装置。

１０．前記第１の制御情報から得られる送信信号の制御量の分散が所定の値より大きい場合には、到来方向推定結果のみを用いて送信制御することを特徴とする請求の範囲第８項に記載の送信ダイバーシチ通信装置。

１１．前記制御手段は、今回受信された第１及び第２の制御情報と、以前に受信した１以上の第１及び第２の制御情報を用いて、フィルタリング処理を行った結果により制御を行うことを特徴とする請求の範囲第１項に記載の送信ダイバーシチ通信装置。

１２．各グループの基準アンテナから送信する信号の送信電力よりも、基準アンテナ以外のアンテナから送信する信号の送信電力を低く設定することを特徴とする請求の範囲第１項に記載の送信ダイバーシチ通信装置。

１３．移動局からの情報を基に送信信号の制御を行う、送信ダイバーシチ方式を適用した基地局を有する送信ダイバーシチ通信装置であって、

複数本ずつのアンテナが複数のグループに分けられ、該グループ内のアンテナは該グループ内のアンテナ間のフェージング相関が高くなるように近接して、各グループのアンテナ群は該グループ間のフェージング相関が低くなるよう互いの距離が大きくなるように設置されたアンテナ手段と、

移動局から送信されてくる該グループ内アンテナの制御に関する送信速度の遅い第１の制御情報と、該各グループのアンテナ群間の制御に関する送信速度の速い第２の制御情報とを受け取って、該アンテナ手段の送信する信号の振幅及び位相を制御する制御手段と、

を備えることを特徴とする送信ダイバーシチ通信装置。

１４．移動局からの情報を基に送信信号の制御を行う、送信ダイバーシチ方式を適用した基地局を有する送信ダイバーシチ通信装置の移動局であって、

複数本ずつのアンテナが複数のグループに分けられ、該グループ内のアンテナは該グループ内のアンテナ間のフェージング相関が高くなるように近接して、各グループのアンテナ群は該グループ間のフェージング相関が低くなるよう互いの距離が大きくなるように設置されたアンテナ手段から送信された信号を受信する受信手段と、

受信した信号がいずれのアンテナから送信された信号であるかを特定するアンテナ特定手段と、

該受信した信号から、該グループ内アンテナの制御に関する第１の制御情報を所定の伝送速度で基地局に送信し、該各グループのアンテナ群間の制御に関する第２の制御情報を該所定の伝送速度より速い速度で基地局に送信する送信手段と、

を備えることを特徴とする移動局。

１５．移動局からの情報を基に送信信号の制御を行う、送信ダイバーシチ方式を適用した基地局を有する送信ダイバーシチ通信方法であって、

複数本ずつのアンテナが複数のグループに分けられ、該グループ内のアンテナは該グループ内のアンテナ間のフェージング相関が高くなるように近接して、各グループのアンテナ群は該グループ間のフェージング相関が低くなるよう互いの距離が大きくなるように設置されたアンテナ群を設けるステップと、

移動局から送信されてくる該グループ内アンテナの制御に関する送信速度の遅い第１の制御情報と、該各グループのアンテナ群間の制御に関する送信速度

の速い第２の制御情報とを受け取って、該アンテナ手段の送信する信号の位相を制御する制御ステップと、
を備えることを特徴とする送信ダイバーシチ通信方法。

１６．移動局からの位相制御情報を基に、基地局の複数アンテナから送信される送信信号の位相制御を行う通信システムであって、

前記基地局の複数のアンテナは、一のアンテナに対して、フェージング相関が高い位置とフェージング相関が低い位置に他のアンテナが配置される位置関係を有し、

前記移動局は、フェージング相関が高い位置に配置されたアンテナに対する位相制御情報は低い頻度で、フェージング相関が低い位置に配置されたアンテナに対する位相制御情報は高い頻度で前記基地局へ送信する
ことを特徴とする通信システム。

１７．移動局からの位相制御情報を基に、基地局の複数のアンテナから送信される送信信号の位相制御を行う通信システムであって、

前記基地局の複数のアンテナは、一のアンテナに対して、フェージング相関が高い位置とフェージング相関が低い位置に他のアンテナが配置される位置関係を有し、

前記移動局は、フェージング相関が高い位置に配置されたアンテナに対する位相制御情報を、フェージング相関が低い位置に配置されたアンテナに対する位相制御情報よりも低い頻度で前記基地局へ送信する
ことを特徴とする通信システム。

１８．移動局からの位相制御情報を基に、基地局の複数のアンテナから送信さ

れる送信信号の位相制御を行う通信システムであって、

前記基地局の複数のアンテナは、一のアンテナに対して、特定のフェージング相関を有する位置に他方のアンテナが配置される位置関係を有し、

前記移動局は、他方のアンテナに対する位相制御情報を、特定のフェージング相関に応じた頻度で前記基地局へ送信する

ことを特徴とする通信システム。

１９．移動局からの位相制御情報を基に、基地局の複数のアンテナから送信される送信信号の位相制御を行う通信システムであって、

前記基地局の複数のアンテナは、一のアンテナに対して、フェージング相関が高い位置に他のアンテナが配置される位置関係を有し、

前記移動局は、他方のアンテナに対する位相制御情報を、低い頻度で前記基地局に送信する

ことを特徴とする通信システム。

２０．移動局からの位相制御情報を基に、基地局の複数のアンテナから送信される送信信号の位相制御を行う通信システムであって、

前記基地局の複数のアンテナは、一のアンテナに対して、フェージング相関が低い位置に他のアンテナが配置される位置関係を有し、

前記移動局は、他方のアンテナに対する位相制御情報を、高い頻度で前記基地局に送信する

ことを特徴とする通信システム。

２１．一のアンテナに対してフェージング相関が高い位置とフェージング相関が低い位置に他のアンテナが配置される位置関係を有する基地局側の複数のア

ンテナから送信される送信信号の位相制御を、移動局からの位相制御情報を基に行う通信システムの移動局であって、

前記複数のアンテナに対する位相制御情報を生成する制御部と、

前記位相制御情報のうち、フェージング相関が高い位置に配置されたアンテナに対する位相制御情報を低い頻度で、フェージング相関が低い位置に配置されたアンテナに対する位相制御情報を高い頻度で前記基地局へ送信する送信部と

を有することを特徴とする移動局。

２２．一のアンテナに対してフェージング相関が高い位置とフェージング相関が低い位置に他のアンテナが配置される位置関係を有する基地局側の複数のアンテナから送信される送信信号の位相制御を、移動局からの位相制御情報を基に行う通信システムの移動局であって、

前記複数のアンテナに対する位相制御情報を生成する制御部と、

前記位相制御情報のうち、フェージング相関が高い位置に配置されたアンテナに対する位相制御情報を、フェージング相関が低い位置に配置されたアンテナに対する位相制御情報よりも低い頻度で前記基地局へ送信する送信部と

を有することを特徴とする移動局。

２３．一のアンテナに対して特定のフェージング相関を有する位置に他方のアンテナが配置される位置関係を有する基地局側の複数のアンテナから送信される送信信号の位相制御を、移動局からの位相制御情報を基に行う通信システムの移動局であって、

前記複数のアンテナに対する位相制御情報を生成する制御部と、

該フェージング相関に応じた頻度で対応するアンテナへの前記位相制御情報

を前記基地局へ送信する送信部と
を有することを特徴とする移動局。

２４．一のアンテナに対してフェージング相関が高い位置に他のアンテナが配置される位置関係を有する基地局側の複数のアンテナから送信される送信信号の位相制御を、移動局からの位相制御情報を基に行う通信システムの移動局であって、

前記複数のアンテナに対する位相制御情報を生成する制御部と、

前記他方のアンテナに対する位相制御情報を、低い頻度で前記基地局に送信する送信部
とを有することを特徴とする移動局。

２５．一のアンテナに対してフェージング相関が低い位置に他のアンテナが配置される位置関係を有する基地局側の複数のアンテナから送信される送信信号の位相制御を、移動局からの位相制御情報を基に行う通信システムの移動局であって、

前記複数のアンテナに対する位相制御情報を生成する制御部と、

前記他方のアンテナに対する位相制御情報を、高い頻度で前記基地局に送信する送信部
とを有することを特徴とする移動局。

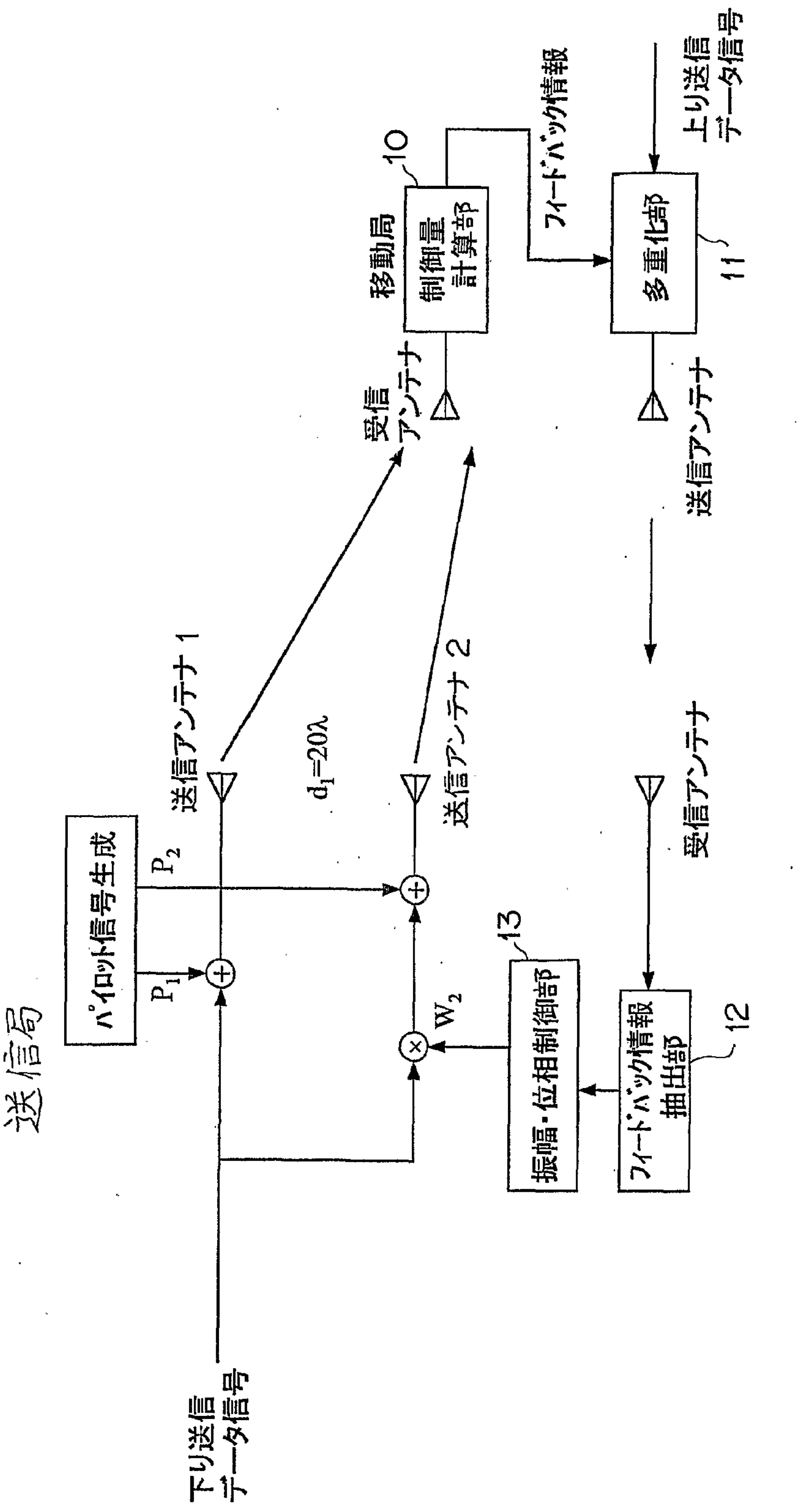

送信局
パイロット信号生成
P_1
P_2
送信アンテナ 1
d_1=20λ
送信アンテナ 2
W_2
振幅・位相制御部
13
フィードバック情報抽出部
12
受信アンテナ
下り送信データ信号
受信アンテナ
移動局
制御量計算部
10
フィードバック情報
多重化部
11
送信アンテナ
上り送信データ信号

図 1

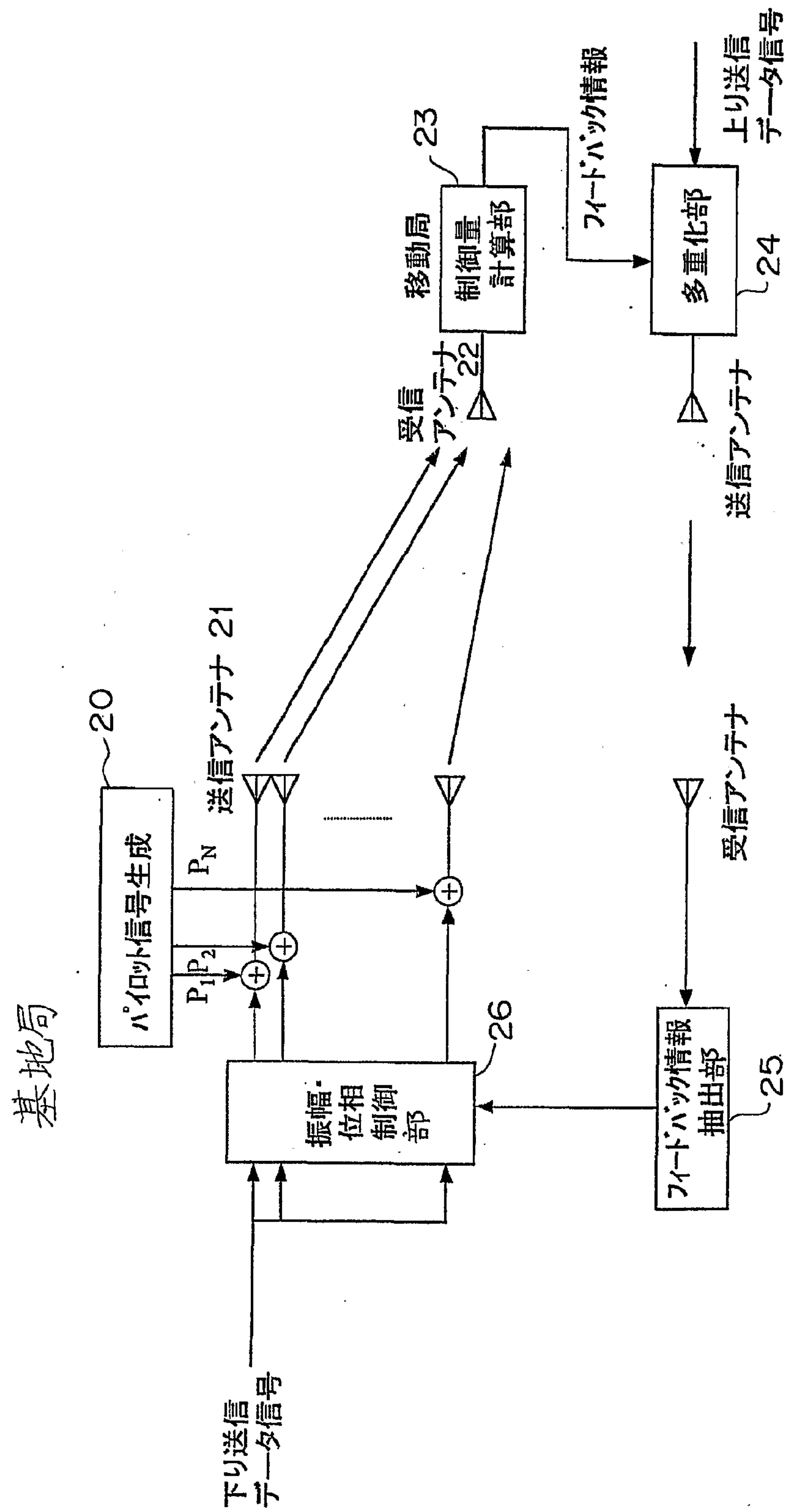

基地局
パイロット信号生成
20
P_1
P_2
P_N
送信アンテナ 21
振幅・位相制御部
26
下り送信データ信号
フィードバック情報抽出部
25
受信アンテナ
移動局
受信アンテナ
22
制御量計算部
23
フィードバック情報
多重化部
24
上り送信データ信号
送信アンテナ

図2

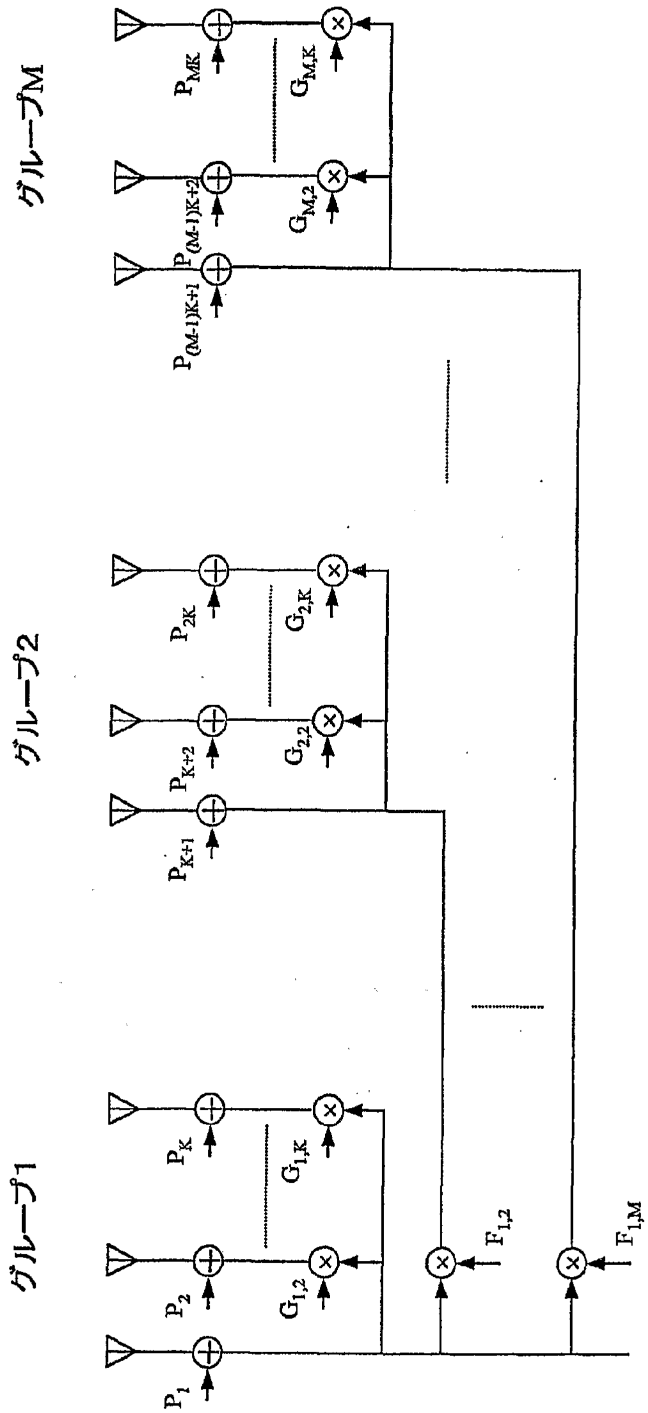

グループ1
グループ2
グループM
P_1
P_2
P_K
P_{K+1}
P_{K+2}
P_{2K}
$P_{(M-1)K+1}$
$P_{(M-1)K+2}$
P_{MK}
$G_{1,2}$
$G_{1,K}$
$G_{2,2}$
$G_{2,K}$
$G_{M,2}$
$G_{M,K}$
$F_{1,2}$
$F_{1,M}$

図 3

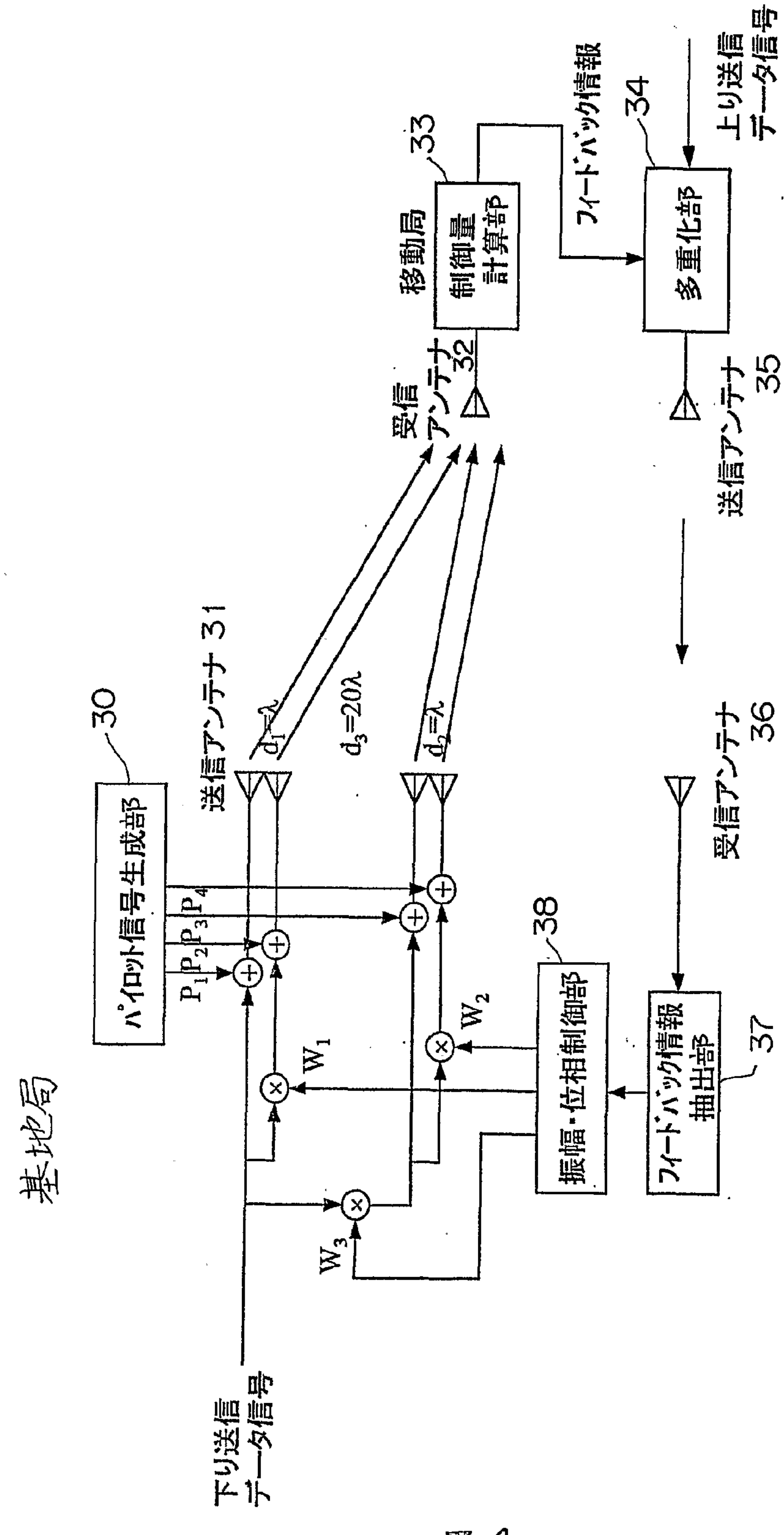
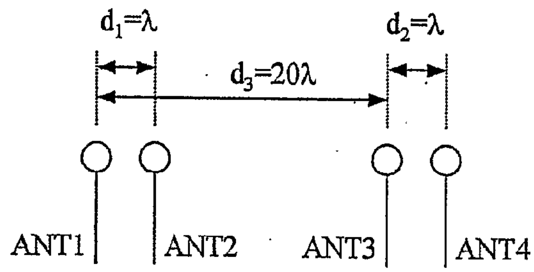

基地局
パイロット信号生成部
30
P_1 P_2 P_3 P_4
送信アンテナ 31
d_1=λ
d_3=20λ
d_2=λ
W_1
W_2
W_3
振幅・位相制御部
38
フィードバック情報抽出部
37
受信アンテナ
36
下り送信データ信号
移動局
受信アンテナ
32
制御量計算部P
33
フィードバック情報
多重化部
34
上り送信データ信号
送信アンテナ
35

図 4

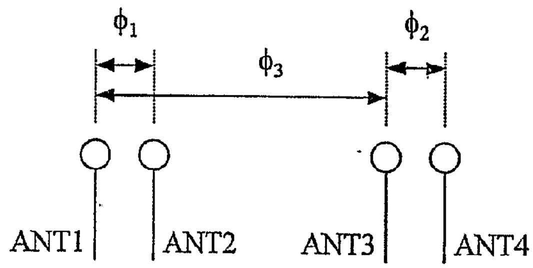
アンテナ1 P_1	A	A	A	A	A	A	A	A	A	A	A	A
アンテナ2 P_2	A	A	-A	-A	A	A	-A	-A	A	A	-A	-A
アンテナ3 P_3	A	-A	A	-A	A	-A	A	-A	A	-A	A	-A
アンテナ4 P_4	A	-A	-A	A	A	-A	-A	A	A	-A	-A	A

A=1+j

図5


$d_1=\lambda$
$d_2=\lambda$
$d_3=20\lambda$
ANT1
ANT2
ANT3
ANT4

(a)


ϕ_1
ϕ_2
ϕ_3
ANT1
ANT2
ANT3
ANT4

(b)

図 6

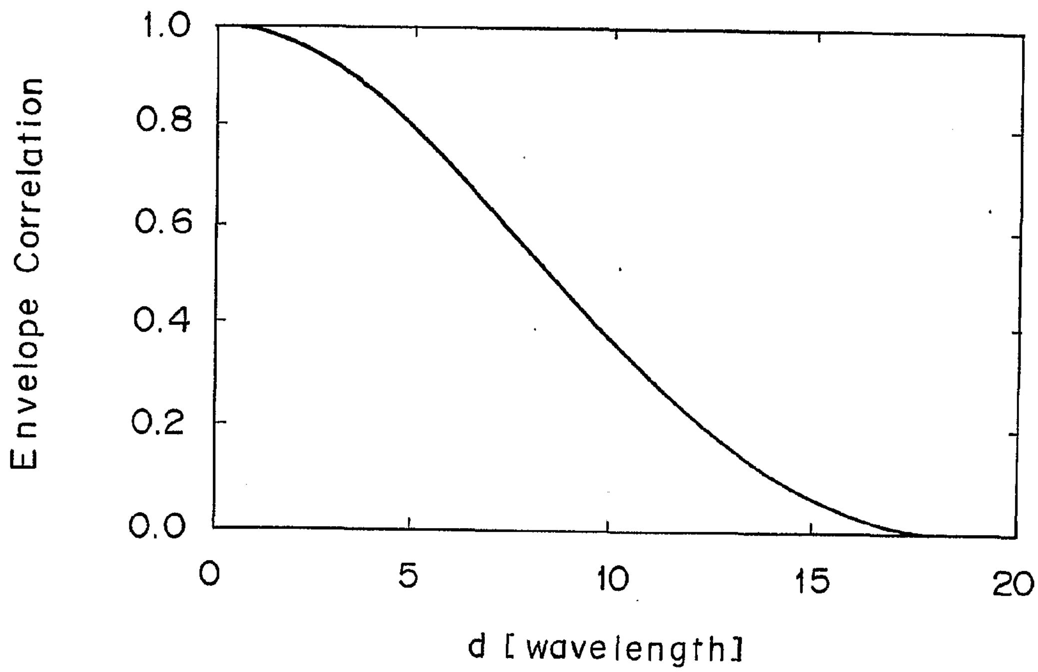

1.0
0.8
0.6
0.4
0.2
0.0
Envelope Correlation
0
5
10
15
20
d [wavelength]

図 7

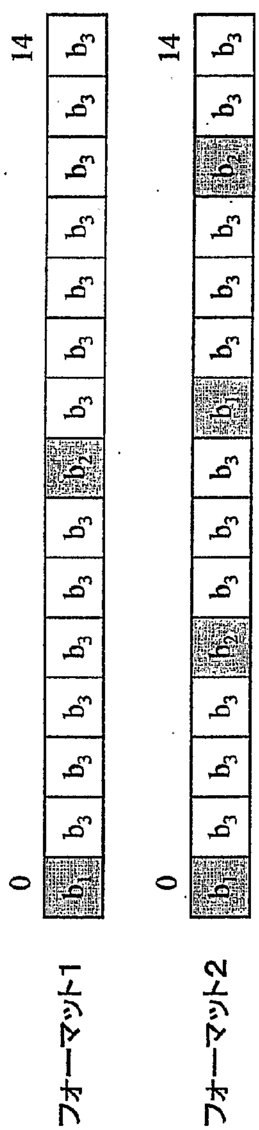

フォーマット1
フォーマット2
0
14
b1 b3 b3 b3 b3 b3 b3 b2 b3 b3 b3 b3 b3 b3 b3
b1 b3 b3 b3 b2 b3 b3 b3 b1 b3 b3 b3 b2 b3 b3

図8

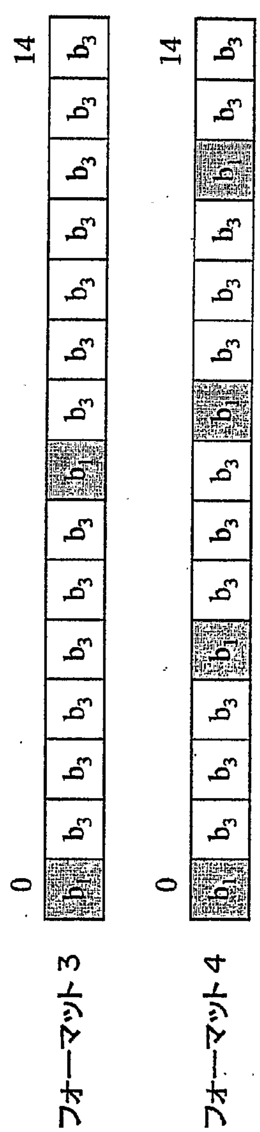

フォーマット3
フォーマット4
0
14
b1 b3 b3 b3 b3 b3 b3 b1 b3 b3 b3 b3 b3 b3 b3
b1 b3 b3 b3 b1 b3 b3 b3 b1 b3 b3 b3 b1 b3 b3

図9

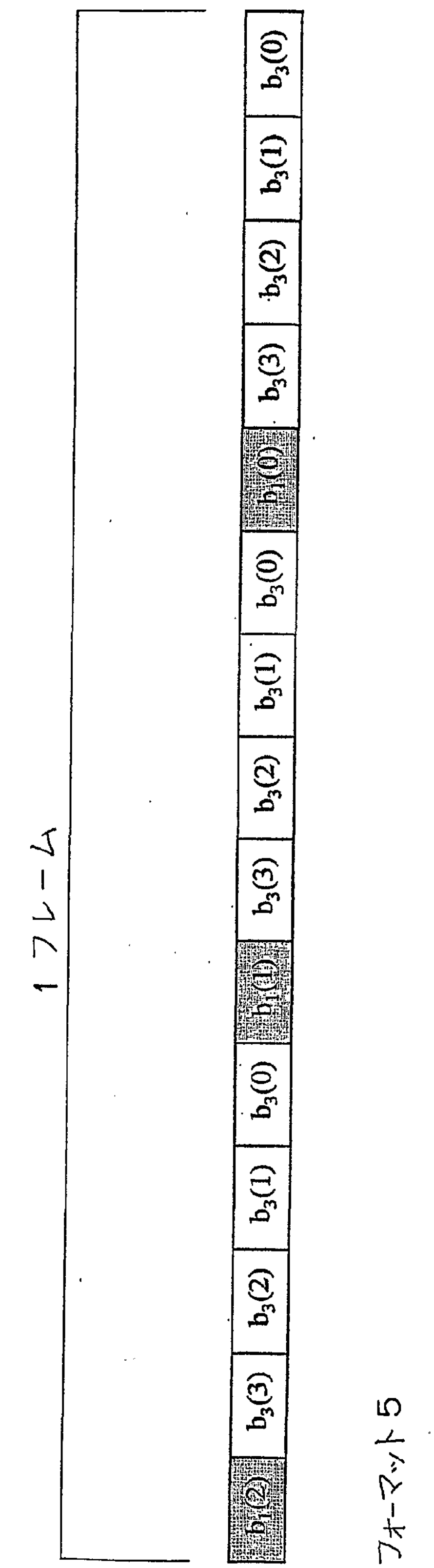

1フレーム
b1(2)
b3(3)
b3(2)
b3(1)
b3(0)
b1(1)
b3(3)
b3(2)
b3(1)
b3(0)
b1(0)
b3(3)
b3(2)
b3(1)
b3(0)
フォーマット5

図10

表1. フィードバックビット($b_3(0)$)

b_3 (0)	アンテナ1振幅	アンテナ2振幅
0	0.2	0.8
1	0.8	0.2

表2. フィードバックビット($b_3(3)$, $b_3(2)$, $b_3(1)$)

$b_3(3)$, $b_3(2)$, $b_3(1)$	アンテナ間位相差(度)
000	180
001	-135
010	-90
011	-45
100	0
101	45
110	90
111	135

表3. フィードバックビット($b_1(2)$, $b_1(1)$, $b_1(0)$)

$b_1(2)$, $b_1(1)$, $b_1(0)$	アンテナ間位相差(度)
000	180
001	-135
010	-90
011	-45
100	0
101	45
110	90
111	135

図 11

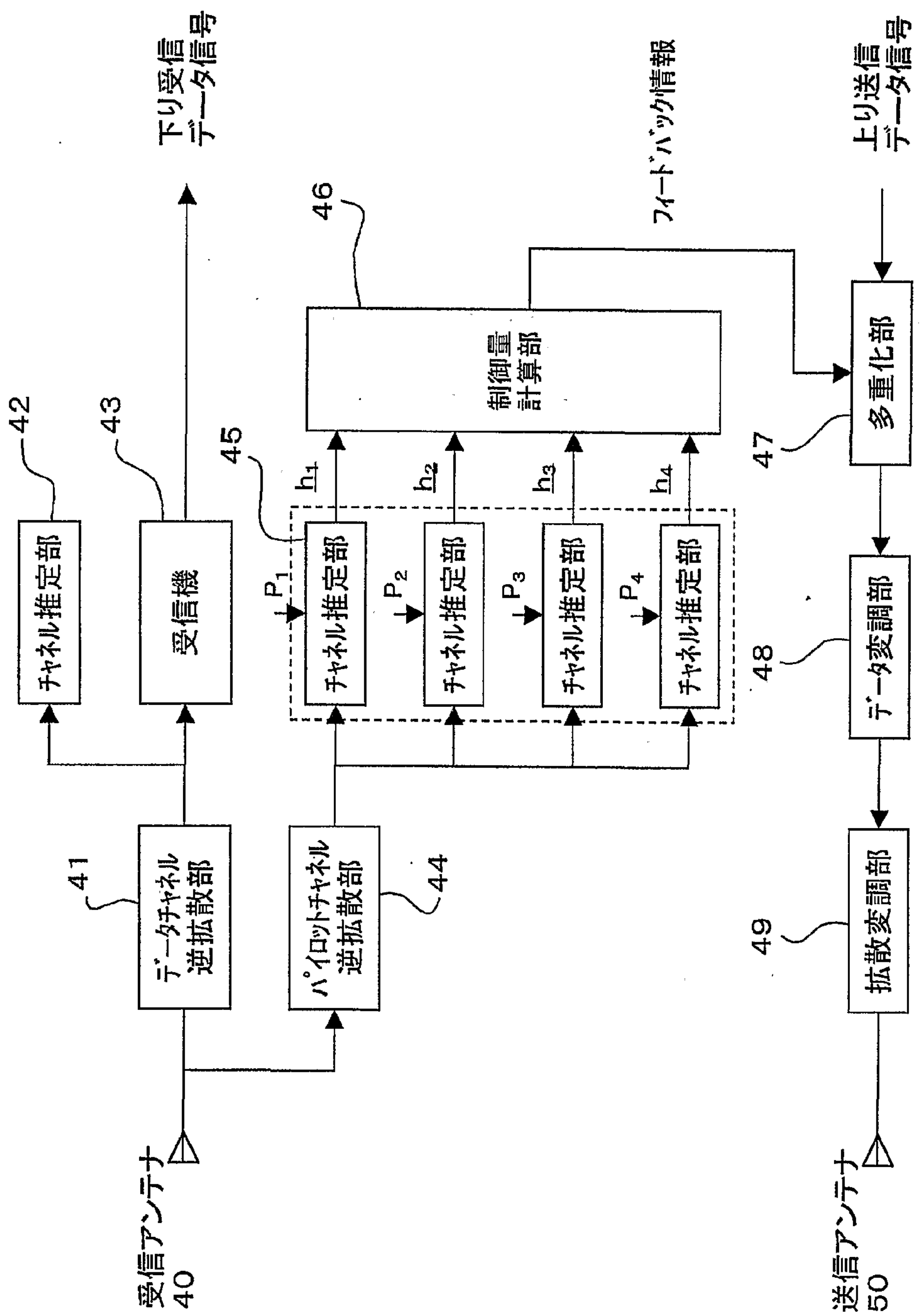

受信アンテナ 40
データチャネル逆拡散部
41
パイロットチャネル逆拡散部
44
チャネル推定部
42
受信機
43
下り受信データ信号
45
P_1
P_2
P_3
P_4
チャネル推定部
チャネル推定部
チャネル推定部
チャネル推定部
h_1
h_2
h_3
h_4
制御量計算部
46
フィードバック情報
上り送信データ信号
多重化部
47
データ変調部
48
拡散変調部
49
送信アンテナ 50

図 12

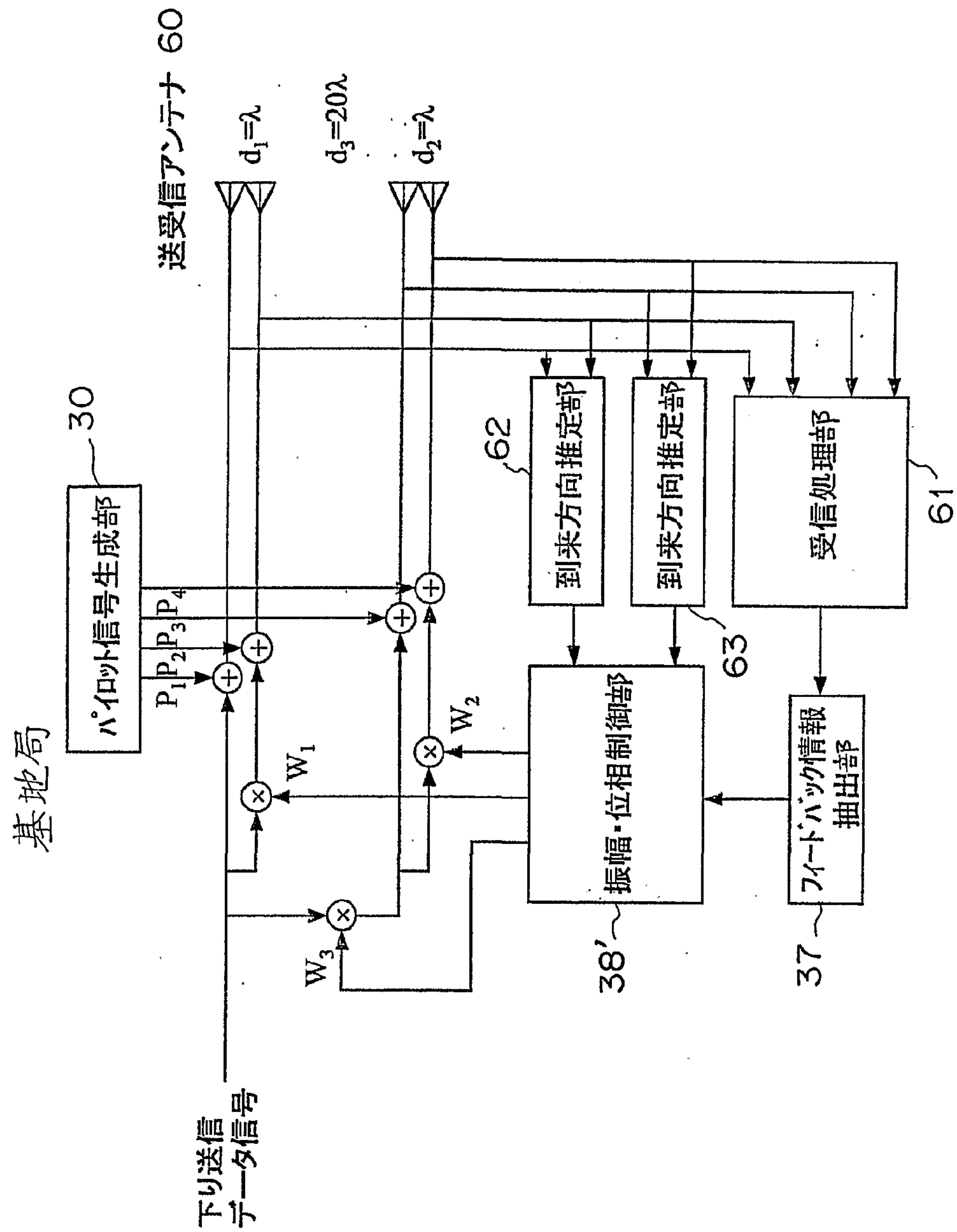

基地局
パイロット信号生成部
30
P1
P2
P3
P4
W1
W2
W3
送受信アンテナ 60
d1=λ
d3=20λ
d2=λ
到来方向推定部
62
到来方向推定部
63
受信処理部
61
振幅・位相制御部
38'
フィードバック情報抽出部
37
下り送信データ信号

図 13

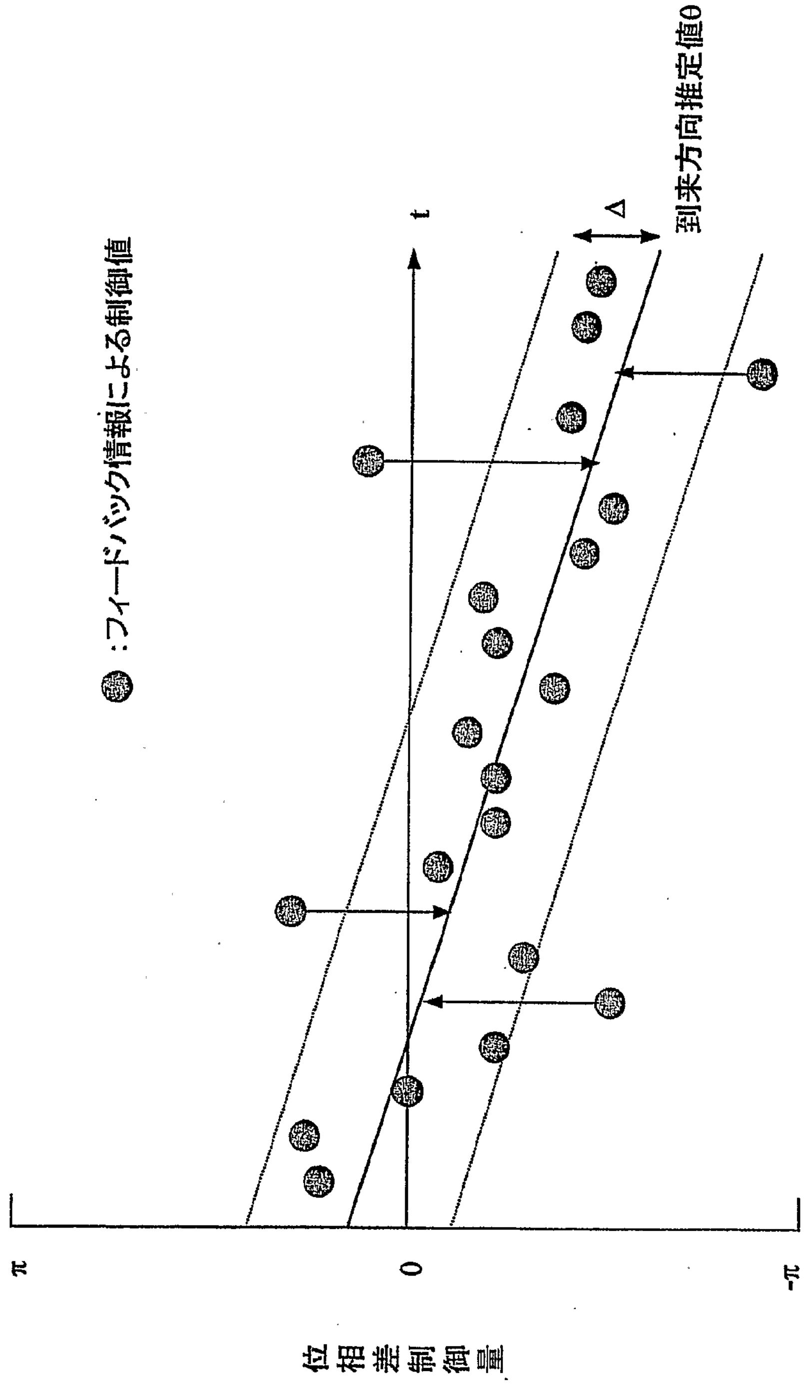


図 14

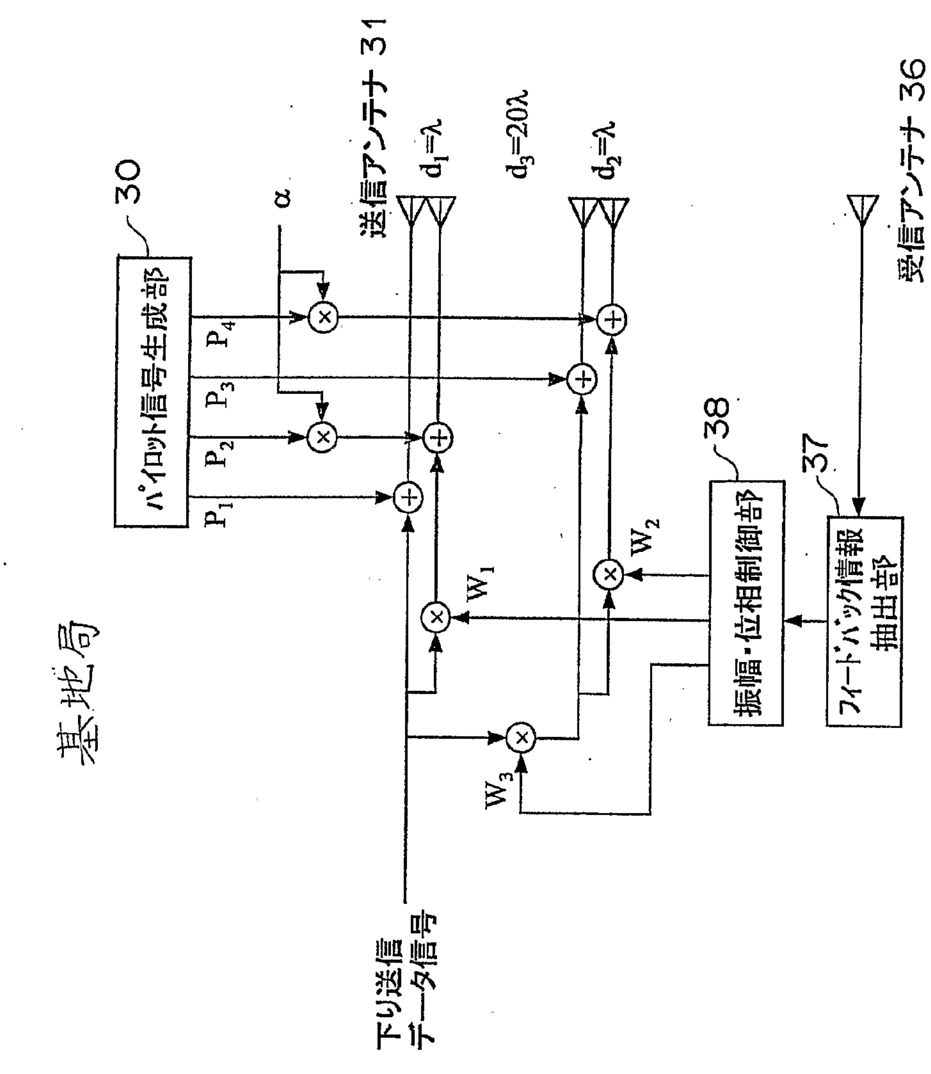

基地局
30
パイロット信号生成部
P1
P2
P3
P4
α
送信アンテナ 31
d1=λ
d3=20λ
d2=λ
受信アンテナ 36
W1
W2
W3
38
振幅・位相制御部
37
フィードバック情報抽出部
下り送信データ信号

図 15

INTERNATIONAL SEARCH REPORT

International application No.
PCT/JP00/05380

A. CLASSIFICATION OF SUBJECT MATTER
Int.Cl^7 H04B 7/06, 7/10, 7/26, H01Q 3/24

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
Int.Cl^7 H01Q 3/00- 3/46, 21/00-25/04
H04B 7/00, 7/02-7/12, 7/24-7/26, 113
H04L 1/02- 1/06, H04Q7/00-7/04

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Jitsuyo Shinan Koho	1922-1996	Toroku Jitsuyo Shinan Koho	1994-2000
Kokai Jitsuyo Shinan Koho	1971-2000	Jitsuyo Shinan Toroku Koho	1996-2000

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y A	JP 58-87928 A (Nippon Telegr. & Teleph. Corp. <NTT>), 25 May, 1983 (25.05.83) (Family: none)	20,25 1-19,21-24
A	JP 10-190537 A (NEC Corporation), 21 July, 1998 (21.07.98) (Family: none)	1-25
A	JP 9-200115 A (Toshiba Corporation), 31 July, 1997 (31.07.97) (Family: none)	1-25

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

* Special categories of cited documents:
"A" document defining the general state of the art which is not considered to be of particular relevance
"E" earlier document but published on or after the international filing date
"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)
"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed
"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
23 October, 2000 (23.10.00)	31 October, 2000 (31.10.00)

Name and mailing address of the ISA/	Authorized officer
Japanese Patent Office	
Facsimile No.	Telephone No.

国際調査報告

国際出願番号　ＰＣＴ／ＪＰ００／０５３８０

Ａ．　発明の属する分野の分類（国際特許分類（ＩＰＣ））
Ｉｎｔ．Ｃｌ⁷　Ｈ０４Ｂ　７／０６，７／１０，７／２６，
Ｈ０１Ｑ　３／２４

Ｂ．　調査を行った分野
調査を行った最小限資料（国際特許分類（ＩＰＣ））
Ｉｎｔ．Ｃｌ⁷　Ｈ０１Ｑ　３／００－　３／４６，２１／００－２５／０４
Ｈ０４Ｂ　７／００，７／０２－７／１２，７／２４－７／２６，１１３
Ｈ０４Ｌ　１／０２－　１／０６，　Ｈ０４Ｑ７／００－７／０４

最小限資料以外の資料で調査を行った分野に含まれるもの
日本国実用新案公報　　　１９２２－１９９６年
日本国公開実用新案公報　１９７１－２０００年
日本国登録実用新案公報　１９９４－２０００年
日本国実用新案登録公報　１９９６－２０００年

国際調査で使用した電子データベース（データベースの名称、調査に使用した用語）

Ｃ．　関連すると認められる文献

引用文献のカテゴリー＊	引用文献名　及び一部の箇所が関連するときは、その関連する箇所の表示	関連する請求の範囲の番号
Ｙ Ａ	ＪＰ，５８－８７９２８，Ａ（日本電信電話公社）２５．５月．１９８３（２５．０５．８３）（ファミリーなし）	20,25 1-19,21-24
Ａ	ＪＰ，１０－１９０５３７，Ａ（日本電気株式会社）２１．７月．１９９８（２１．０７．９８）（ファミリーなし）	1-25
Ａ	ＪＰ，９－２００１１５，Ａ（株式会社東芝）３１．７月．１９９７（３１．０７．９７）（ファミリーなし）	1-25

☐ Ｃ欄の続きにも文献が列挙されている。　☐ パテントファミリーに関する別紙を参照。

＊　引用文献のカテゴリー
「Ａ」特に関連のある文献ではなく、一般的技術水準を示すもの
「Ｅ」国際出願日前の出願または特許であるが、国際出願日以後に公表されたもの
「Ｌ」優先権主張に疑義を提起する文献又は他の文献の発行日若しくは他の特別な理由を確立するために引用する文献（理由を付す）
「Ｏ」口頭による開示、使用、展示等に言及する文献
「Ｐ」国際出願日前で、かつ優先権の主張の基礎となる出願の日の後に公表された文献
「Ｔ」国際出願日又は優先日後に公表された文献であって出願と矛盾するものではなく、発明の原理又は理論の理解のために引用するもの
「Ｘ」特に関連のある文献であって、当該文献のみで発明の新規性又は進歩性がないと考えられるもの
「Ｙ」特に関連のある文献であって、当該文献と他の１以上の文献との、当業者にとって自明である組合せによって進歩性がないと考えられるもの
「＆」同一パテントファミリー文献

国際調査を完了した日 ２３．１０．００	国際調査報告の発送日 31.10.00
国際調査機関の名称及びあて先 日本国特許庁（ＩＳＡ／ＪＰ） 郵便番号１００－８９１５ 東京都千代田区霞が関三丁目４番３号	特許庁審査官（権限のある職員）　５Ｊ　９６５４ 徳田　賢二 電話番号　０３－３５８１－１１０１　内線　３５３６

PCT

WORLD INTELLECTUAL PROPERTY ORGANIZATION
International Bureau

INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(51) International Patent Classification 6 : H04J 9/00, H04L 5/04, H04Q 7/00, H04B 1/06, 7/00, 7/216, 7/208 — A1

(11) International Publication Number: WO 00/11823

(43) International Publication Date: 2 March 2000 (02.03.00)

(21) International Application Number: PCT/US98/17063

(22) International Filing Date: 18 August 1998 (18.08.98)

(71) Applicant *(for all designated States except US)*: RADIX TECHNOLOGIES, INC. [US/US]; 329 N. Bernardo Avenue, Mountain View, CA 94043 (US).

(72) Inventor; and
(75) Inventor/Applicant *(for US only)*: AGEE, Brian, G. [US/US]; 1596 Wawona Drive, San Jose, CA 95125 (US).

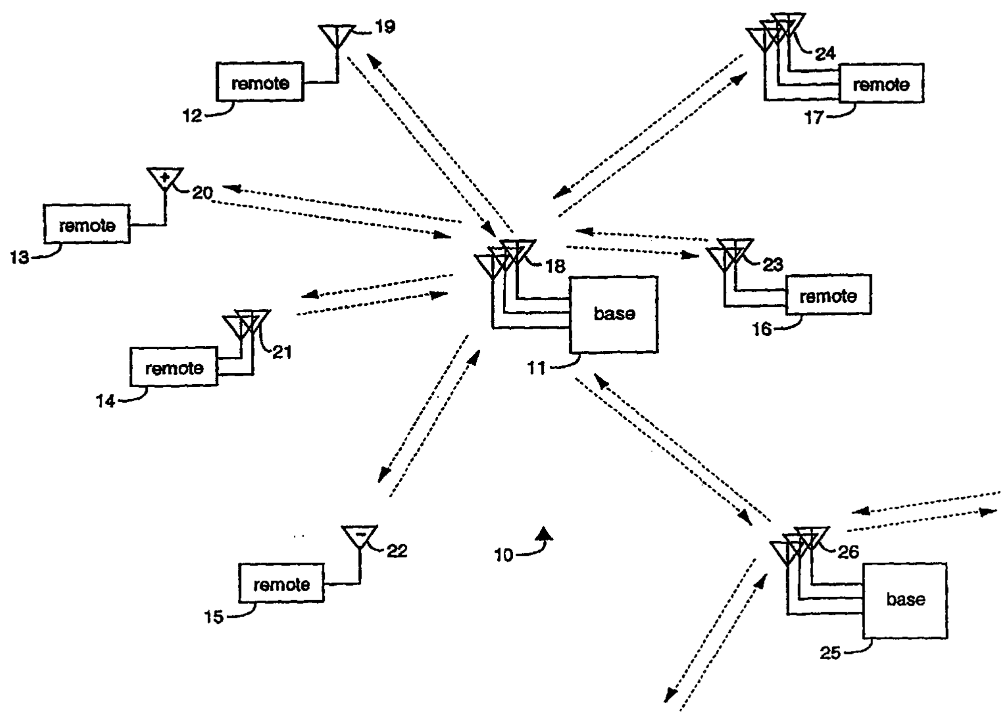
(74) Agents: JAKOPIN, David, A. et al.; Cushman Darby & Cushman, IP Group of Pillsbury Madison & Sutro, LLP, 1100 New York Avenue N.W., Washington, DC 20005 (US).

(81) Designated States: AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, CA, CH, CN, CU, CZ, DE, DK, EE, ES, FI, GB, GE, GH, GM, HR, HU, ID, IL, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MD, MG, MK, MN, MW, MX, NO, NZ, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TR, TT, UA, UG, US, UZ, VN, YU, ZW, ARIPO patent (GH, GM, KE, LS, MW, SD, SZ, UG, ZW), Eurasian patent (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European patent (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE), OAPI patent (BF, BJ, CF, CG, CI, CM, GA, GN, GW, ML, MR, NE, SN, TD, TG).

Published
With international search report.

(54) Title: STACKED–CARRIER DISCRETE MULTIPLE TONE COMMUNICATION TECHNOLOGY

(57) Abstract



A "stacked–carrier" spread spectrum communication system (10) based on frequency domain spreading that multiplies a time–domain representation of a baseband signal by a set of superimposed, or stacked, complex sinusoid carrier waves. In a preferred embodiment (10), the spreading energizes the bins of a large fast Fourier transform (FFT). This provides a considerable savings in computational complexity for moderate output FFT sizes. Point–to–Multipoint and multipoint–to–multipoint (nodeless) network topologies are possible. A code–nulling method is included for interference cancellation and enhanced signal separation by exploiting the spectral diversity of the various sources (11). The basic system (10) may be extended to include multi–element antenna array (26/18) nulling methods also for interference cancellation and enhanced signal separation using spatial separation. Such methods permit directive and retrodirective transmission systems that adapt or can be adapted to the radio environment. Such systems are compatible with bandwidth–on–demand and higher–order modulation formats and use advanced (maximum–SINR) despreader adaptation algorithms.

STACKED-CARRIER DISCRETE MULTIPLE TONE COMMUNICATION TECHNOLOGY

BACKGROUND OF THE INVENTION

Field of the Invention

This invention relates generally to radio communications and more specifically to communication technologies for multiple access in difficult and hostile environments combined with dynamic environment changes.

Description of the Prior Art

The communication technology developed in the 1940's during World War II included "frequency diversity communication" or "stacked carrier communications" to aid high frequency (HF) band traffic. J. Proakis refers to frequency diversity communication technology in, Digital Communications, McGraw-Hill, 1989, see, sections 7.4 to 7.7. Diversity techniques are said by Proakis to be based on the notion that errors occur in the reception of largely attenuated channels, e.g., channels in deep fade. Supplying the receiver with several duplicates of the original signal, but over channels that fade independent from one another, has the potential of securing continuous communication except during the unlikely event that all the duplicate channels fade out together. Such probability can be estimated.

Frequency diversity is one of many diversity methods. The same modulation is carried by several carrier channels separated by nominally the coherence bandwidth of each respective channel. In time diversity, the same information is transmitted over different time slots. Multiple antennas can be used in a diversity scheme. Several receiving antennas can be used to receive the signals sent from a single transmitting antenna. For best effect, the receiving antennas are spaced enough apart to vary different multipath interference amongst the group. A separation of nominally ten wavelengths is generally needed to observe independent signal fading.

A signal having a bandwidth much greater than the coherence bandwidth of the channel can be used in a more sophisticated diversity scheme. Such a signal with a bandwidth W will resolve the multipath components and provide the receiver with several independently fading signal paths.

Other prior art diversity schemes have included angle-of-arrival or spatial diversity and polarization diversity.

When a bandwidth W much greater than the coherence bandwidth of each respective channel is available to a user, the channel can be subdivided into a number of frequency division

multiplexed sub-channels having a mutual separation in center frequencies of at least the coherence bandwidth of each respective channel. The same signal can then be transmitted over the frequency-division multiplex sub-channels to establish frequency diversity operation. The same result can be achieved by using a wideband binary signal that covers the bandwidth W.

G. K. Kaleh describes such in an article, "Frequency-Diversity Spread-Spectrum Communication System to Counter Band-limited Gaussian Interference," IEEE Transactions on Communications, Sept. 1994. Here a secure setup is outlined that can operate in deliberately hostile signal environments.

J. Proakis describes frequency diversity spread spectrum and multiple access concepts in chapter eight, "Spread Spectrum Signals for Digital Communication," supra. Diversity transmission combined with frequency-hopping spread spectrum is detailed for protection against multipath fading and partial-band jamming.

Retro-directivity was proposed and used as early as 1959 to adapt a multi-element antenna array to provide identical spatial gain patterns during transmission and reception operations. See, R. Monzingo, T. Miller, Introduction to Adaptive Arrays, Wiley Interscience Publications, 1980; L. Van Atta, "Electromagnetic Reflection," U.S. Patent 2,908,002, 1959; and B. Glance, P. Henry, "High Capacity Mobile Radio System," U.S. Patent 4,383,332, May 10, 1983, for a discussion of such techniques. TDD systems provide an effective means for implementing retro-directive antenna arrays, e.g., by minimizing the channel variation between the reception and transmission paths.

Summary of the Invention

It is therefore an object of this invention to provide a radio communication system for spreading data over widely separated frequency bands manifesting differences in channel distortion without physically spreading signals between intervening frequencies, as is required with direct-sequence spread spectrum.

It is another object of the invention to provide a radio communication system for communication under strong narrow-band interference, e.g., conventional cellular signal waveforms, by turning off affected frequency channels at a receiver's despreader.

It is an object of the invention to provide a radio communication system with simple equalization of linear channel multipath distortion.

It is an object of the invention to provide a radio communication system that is compatible with discrete multitone and orthogonal frequency-division multiplex-like channelization techniques. And that is compatible with time-packetized discrete multitone and orthogonal

frequency-division multiplex-like modulation/demodulation techniques for frequency channelization and inverse channelization.

It is another object of this invention to provide a radio communication system that is compatible with time-division duplex systems where the stacked-carrier spread spectrum modulation format is packetized, e.g., if the stacked-carrier spread spectrum signal is generated using discrete multiple tone and/or orthogonal frequency-division multiplex-like based frequency channelizers and inverse channelizers.

It is an object of the invention to provide a radio communication system for frequency-division multiple-access like multiple access capability.

It is an object of the invention to provide a radio communication system for code- division multiple-access like capability in a stacked carrier multiple access arrangement.

It is an object of the invention to provide a radio communication system compatible with high-order digital modulations.

It is an object of the invention to provide a radio communication system for bandwidth-on-demand flexible data rate connections.

It is an object of the invention to provide a radio communication system for space-division multiple-access like multiple access, interference excision, and channel equalization capability in a code nulling application.

It is an object of the invention to provide a radio communication system for use with adaptive antenna arrays by spatially extending a spreading code to spread data using independent complex gains on each spatial channel, or antenna beam, to control the channel-bandwidth array dispersion.

It is an object of the invention to provide a radio communication system compatible with advanced array adaptation techniques, e.g., non-blind pilot-directed, blind data-directed, and other techniques that take advantage of underlying properties of the baseband data, channel structure, or stacked carrier spreading format.

It is an object of the invention to provide a radio communication system compatible with retrodirective communication techniques.

It is an object of the invention to provide a radio communication system for back-compatibility with conventional code-division multiple access, data activation systems.

Briefly, an embodiment of the invention comprises a "stacked-carrier" spread spectrum communication system wherein the spreading is done in the frequency domain by multiplying a time-domain representation of a baseband signal by a set of superimposed, or stacked, complex sinusoid carrier waves. In practice, the spreading is done by simply energizing the

bins of a large fast Fourier transform (FFT). This provides a considerable savings in computational complexity for moderate output FFT sizes. A Kaiser-Bessel window, e.g., with ß = 9, can be used to "fill out" the space between the tones without subjecting those tones to unacceptable interference from adjacent tones, e.g., inter-tone interference. In particular, a high value of ß provides acceptable interference between adjacent tones and extremely low inter-ference between farther out tones. This basic technology is then combined with time division duplex, code-division multiple access, space-division multiple access, frequency-division multiple access, adaptive antenna array and interference cancellation techniques.

An advantage of the invention is that a radio communication method is provided that spreads data over widely separated frequency bands for spectral diversity. This provides an efficient way to take advantage of frequency diversity, especially in applications where the bands are very widely separated.

An advantage of the invention is that a radio communication method is provided that communicates even under strong narrow-band interference. So a stacked-carrier spread spectrum (SCSS) link can be maintained in the presence of strong narrow-band frequency-division multiple access (FDMA) and time-division multiple access (TDMA) cellular radio signals, as in cellular overlay applications. It also allows such a link to be maintained in the presence of spurious interference due to harmonics from out-of-band signals.

An advantage of the invention is that a radio communication method is provided that permits simple equalization of linear channel distortion, and allows stationary, or quasi-stationary, linear channel distortion to be approximated as a multiplicative effect on the transmit spreading code. It further allows the channel equalization operation to be subsumed into the despreading or spreading operation without additional filtering operations, apart from removal of intrapacket Doppler spread. The basic technique equalizes multipath dispersion commensurate with the bandwidth of the baseband, pre-spread, message signal. This multipath equalization operation can be extremely simple if the bandwidth of the message signal is low. If the bandwidth of the pre-spread message signal is sufficiently low, e.g., the correlation width or inverse bandwidth of the prespread message signal is a large multiple of the largest multipath delay in the transmission channel, this equalization operation reduces to a complex multiply operation that is automatically subsumed into the adaptive despreading operation. This is in contrast to conventional CDMA systems, which require additional equalization operations unless the correlation width of the spread signal is a large multiple of the largest multipath delay in the transmission channel.

Another advantage of the invention is that a radio communication method is provided that is compatible with discrete multiple tone and orthogonal frequency-division multiplex-like frequency channelization techniques. This allows stationary and linear channel distortion to be modeled as an exactly multiplicative effect on the transmit spreading code.

An advantage of the invention is that a radio communication method is provided that is compatible with time-division duplex systems. So time-division duplex communication formats can be used where the stacked-carrier spread spectrum modulation format is packetized, e.g., if the stacked-carrier spread spectrum signal is generated using discrete multiple tone and/or orthogonal frequency-division multiplex-like based frequency channelizers and inverse channelizers. A "local" estimation of the transmit channel at either end of the communication link is made possible, greatly simplifying the implementation of channel preemphasis, transmit-site channel equalization topologies and retrodirective transmission techniques.

An advantage of the invention is that a radio communication method is provided with a code-division multiple-access type of multiple access capability, e.g., the stacked carrier multiple access technique. Point-to-multipoint communication links are implemented by transmitting signals over the same subset of frequency channels, using linearly interdependent, (orthogonal or non-orthogonal) sets of spreading gains to separate the signals at the despreader. Since the spreading codes can be non-orthogonal, a chief advantage of this invention when used in conjunction with code nulling techniques, is that the use of non-orthogonal spreading codes is possible.

An advantage of the invention is that a radio communication method is provided that is compatible with "bandwidth-on-demand" flexible data rate techniques. The data rate supplied on a given link can be increased or decreased in small increments by transmitting primitives to a single user over multiple time, frequency, or stacked carrier channels. The data rate is then adjusted with no increase in bandwidth if the data rate increases using multiple stacked carrier channels.

An advantage of the invention is that a radio communication method is provided that is compatible with high-order digital modulations. It is compatible with arbitrary M_{ary} digital baseband modulation formats and allows capacity improvement through transmission of higher numbers of bits/symbol on each frequency channel. The reuse is improved and "load balancing" in multicell communication networks can be included by varying the bits per symbol on each primitive.

An advantage of the invention is that a radio communication method is provided that has space-division multiple-access, interference excision, and channel equalization capability, e.g.,

in the code nulling technique. Such space-division multiple-access like code nulling techniques, in optimal and quasi-optimal linear interference cancellation and signal extraction techniques, are useful to separate stacked-carrier spread spectrum signals at the despreader, based on the frequency diversity or spectral diversity of the signals. Interference excision against in-cell stacked-carrier spread spectrum signals are thereby provided, as well as the elimination of out-of-cell interferers, e.g., the reuse enhancement capability. This then allows the most effective use of code nulling, which is generally applicable to a wide range of spreading formats. In particular, such provides a factor of two capacity improvement over code-nulling techniques developed for use with modulation-on-symbol direct-sequence spread spectrum formats where the spreading gain repeats once every underlying message symbol.

An advantage of the invention is that a radio communication method is provided that can be used with adaptive antenna arrays.

An advantage of the invention is that a radio communication method is provided that is compatible with advanced array adaptation techniques and thereby separates signals based on spatial diversity, frequency spectrum diversity, polarization diversity, and combinations of spatial/spectral/polarization diversity.

An advantage of the invention is that a radio communication method is provided that is compatible with retrodirective communication techniques. This enables a straightforward extension of spatial retro-directivity technique to stacked-carrier spread spectrum systems including single antennas or antenna arrays. And allows concentration of most complex operations at the base station in point-to-multipoint communication links, greatly reducing the cost of the overall system.

A still further advantage is that a radio communication method is provided that is backward compatible with conventional code-division multiple access, data activation techniques.

These and other objects and advantages of the invention will no doubt become obvious to those of ordinary skill in the art after having read the following detailed description of the preferred embodiment which is illustrated in the various figures.

In the Drawings

Fig. 1 is a block diagram of a communication system embodiment of the invention wherein several remote mobile units are distributed in space about one or more central base stations;

Fig. 2A is a block diagram representing an embodiment of the invention wherein a stacked-carrier spread-spectrum transmitter bank is connected to an antenna array for a point-to-point transmitter, and another antenna array is connected to a stacked-carrier spread-spectrum receiver bank 5 for a point-to-point receiver;

Fig. 2B is a block diagram representing another embodiment of the invention wherein a stacked-carrier multiple-access transmitter bank is connected to an antenna array for a network transmitter, and another antenna array is connected to a stacked-carrier multiple-access receiver bank for a network receiver;

Fig. 3A is a block diagram representing another embodiment of the invention wherein a stacked-carrier spread-spectrum transmitter is connected to a time-division duplexer for a point-to-point transmitter, and another time-division duplexer is connected to a stacked-carrier spread-spectrum receiver for a point-to-point receiver;

Fig. 3B is a block diagram representing another embodiment of the invention wherein a stacked-carrier multiple-access transmitter is connected to a time-division duplexer for a network transmitter, and another time-division duplexer is connected to a stacked-carrier multiple-access receiver for a network receiver;

Fig. 4A is a block diagram representing another embodiment of the invention wherein a stacked-carrier spread-spectrum transmitter is connected to a code nuller for a point-to-point transmitter, and another code nuller is connected to a stacked-carrier spread-spectrum receiver for a point-to-point receiver;

Fig. 4B is a block diagram representing another embodiment of the invention wherein a stacked-carrier multiple-access transmitter is connected to a code nuller for a network transmitter, and another code nuller is connected to a stacked-carrier multiple-access receiver for a network receiver;

Fig. 5A is a block diagram representing another embodiment of the invention wherein a stacked-carrier spread-spectrum transmitter is connected to a widely separated frequency channelizer for a point-to-point transmitter, and another widely separated frequency channelizer is connected to a stacked-carrier spread-spectrum receiver for a point-to-point receiver;

Fig. 5B is a block diagram representing another embodiment of the invention wherein a stacked-carrier multiple-access transmitter is connected to a widely separated frequency channelizer for a network transmitter, and another widely separated frequency channelizer is connected to a stacked-carrier multiple-access receiver for a network receiver;

Fig. 6A is a block diagram representing another embodiment of the invention wherein a stacked-carrier spread-spectrum transmitter bank is connected to a synchronized time-division duplexer bank that is connected to both an antenna array and a stacked-carrier spread-spectrum receiver bank with a retro-adaptor in control of the stacked-carrier spread-spectrum transmitter bank for a point-to-point transceiver system;

Fig. 6B is a block diagram representing another embodiment of the present invention wherein a stacked-carrier multiple access transmitter bank is connected to a synchronized time-division duplexer that is connected to both an antenna array and a stacked-carrier multiple access receiver bank with a retro-adaptor in control of the stacked-carrier multiple access transmitter bank for a network system;

Fig. 7A is a functional block diagram of a stacked-carrier spread-spectrum transmitter similar to those included in Figs. 2A, 3A, 4A, 5A, and 6A;

Fig. 7B is a functional block diagram of a stacked-carrier spread-spectrum receiver similar to those included in Figs. 2A, 3A, 4A, 5A, and 6A;

Fig. 8 is a block diagram of the base station included in the system of Fig. 1 and showing the possibility of antenna arrays that allow spatial discrimination amongst the members of the communication system. Each functional transmitter and receiver line is represented as comprising many channels in support of the basic stacked carrier spread spectrum communication medium;

Fig. 9 is a block diagram of a typical remote unit included in the system of Fig. 1 and showing adaptive channel equalization and preemphasis functions in support of the basic stacked carrier spread spectrum communication medium;

Fig. 10 is a block diagram of a multi-element T/R module that includes a plurality of individual T/R modules, one for each antenna. The system complexity can be scaled up or down with the number of antennas. Spatial processing occurs after the analog-to-digital conversion (ADC) process during the reception operation, and before the digital-to-analog conversion (DAC) operation during the transmission operation. All of the spatial as well as spectral spreading operations are performed on the digital data. All of the key frequency and clock references in the system are derived from a common clock, such as a GPS clock. A mechanism is shown for module calibration, which is necessary for accurate retro-directivity in the TDD system;

Fig. 11 is a block diagram of a stacked carrier spread spectrum modulator wherein baseband data is replicated over K_{spread} separate spreading cells multiplied by a separate scalar that is passed to a time-multiplexer for combination into a complex data vector;

Fig. 12 is a block diagram of a stacked carrier spread spectrum despreader in an all-digital fully-adaptive implementation. The despreader includes several channels for processing each of the tones in a stacked carrier spread spectrum carrier medium;

Fig. 13 diagrams an exemplary BPSK multitone having a data length of six, a spreading factor K_{spread} of four and a separation between each group of two. Each group cell, gl-g4, is

represented as having an independent amplitude that can be manipulated by channel equalization and preemphasis to combat interference and other problems;

Fig. 14 diagrams a "SCORE" processor used to restore a received signal x(t) from an antenna array. The processor controls include control filter h(t), frequency-shift value alpha and a conjugation control (*);

Fig. 15 is a dataflow diagram representing a code-gated SCORE despreading operation with gating over two cell subsets;

Fig. 16 is a dataflow diagram representing a code-gated SCORE spreading operation with gating over two cell subsets, and demonstrates a symmetry with that of Fig. 15;

Fig. 17 is a time-frequency format for a time-division duplex communication system embodiment of the invention;

Fig. 18 is an active tone format of a basic DMT modem;

Fig. 19 is a dataflow diagram representing a transmit/receive calibration method;

Fig. 20 is a diagram of an integrated single-antenna T/R and discrete multiple tones (DMT) modem to implement a DMT based stacked-carrier multiple-access (SCMA) system embodiment of the invention;

Fig. 21 is a general example of a single-link code-gated cross-SCORE spreader embodiment of the invention;

Fig. 22 is a dataflow diagram representing a single-link code-gated cross-SCORE despreading operation with K_{spread} cell subsets;

Fig. 23 is a dataflow diagram representing a single-link cross-SCORE algorithm with N_{frame} packets/adapt frame;

Fig. 24 is a dataflow diagram representing a single adapt frame autocorrelation statistics computation;

Fig. 25 is a dataflow diagram representing a cross-SCORE eigenequation with K_{spread} cell subsets;

Fig. 26 is a dataflow diagram representing a code key generator with $K_{part} < K_{spread}$ cell subsets;

Fig. 27 is a dataflow diagram representing an equivalent code key applicator with K_{part} K_{spread} cell subsets;

Fig. 28 is a dataflow diagram representing a cross-SCORE eigenequation with K_{part} subsets;

Fig. 29 is a dataflow diagram representing a cross-SCORE eigenequation with two cell subsets;

Fig. 30 is a dataflow diagram representing a multi-link code-gated cross-SCORE spreader embodiment of the invention;

Fig. 31 is a dataflow diagram representing a single-link code-gated auto-SCORE spreading operation with gating over frequency and two cell subsets in an embodiment of the invention;

Fig. 32 is a dataflow diagram representing a single-link code-gated auto-SCORE despreading operation with gating over frequency and two cell subsets;

Fig. 33 is a dataflow diagram representing an auto-SCORE eigenequation with gating over frequency and two cell subsets;

Fig. 34 is a dataflow diagram representing a single-link code-gated auto-SCORe spreading with gating over time and a rate one-half of redundancy gate; and

Fig. 35 is a dataflow diagram representing a single-link code-gated auto-SCORE despreading with gating over time and a rate one-half of redundancy gate.

Detailed Description of the Preferred Embodiment

Fig. 1 illustrates a communication system embodiment of the invention, referred to herein by the general reference numeral 10. The system 10 comprises a base station 11 in two-way radio communication with a plurality of remote units 12-17. The positions of the remote units 1217 around the bane station 11 in Fig. 1 represents the variety of different positions in three-dimensional space that can be assumed by all, or by one or more remotes at various points in time. The base station 11 has a multi-element antenna 18. Each remote unit 12-17 has a corresponding antenna 19-24, some or all of which may comprise multi-element antennas, e.g., antennas 21, 23 and 24. The antennas 18-24 represent alternatives that range from a single physical antenna connected to a transceiver, to separate transmit and receive antennas, and to arrays of antennas that each express differential spatial signal sensitivities. Moreover, some or all of antennas 18-24 may be polarization diverse. That is, certain of antennas 18-24 may be positive sense polarized (e.g., antenna 20) and certain may be negative sense polarized (e.g., antenna 22). The "positive/negative" polarization sense may be based on "horizontal/vertical" linear polarization, "clockwise/counterclockwise" circular polarization, "slant 45/135" polarization, etc. True noise invades the system 10 equally from all directions and sources of interference are typically defined by their signals which arrive from particular directions. Multipath signals between the base 11 and the remote units 12-17 represent a type of interference that can cause channel fade and other problems.

The system 10 can also include multipoint-to-multipoint and point-to-point network topologies, as represented by a second base station 25 with a multi-element antenna 26. The

multipoint-to-multipoint network is a superset of that shown in Fig. 1, and is useful in cell systems where adjacent call interface needs to be controlled. Each base or remote transceiver in the network can have arbitrarily different numbers of antenna elements and spreading factors, e.g., they may be spread over different numbers of frequency cells. Spatially localized interference can arrive from other stacked carrier networks and cells within the network and from other interferers, e.g., jammers, or FDMA signals that the network is being layered over. True noise can invade the system equally or unequally from all directions, wherein "equally" can imply isotropic noise.

The basic means of radio communication in the system 10 is, by what is called here, "stacked-carrier spread spectrum" (SCSS), wherein discrete multiple tones (DMT) having a substantial frequency diversity are simultaneously transmitted by the base station 11 and by each remote unit 12-17 to the other. One baseband data symbol is spread spectrum modulated on each set of discrete multitone transmissions from a single unit 11-17. Accurate data recovery can be accomplished by the intended receiver even though some of the individual channels carrying information on a discrete tone may have faded out or been interfered with too severely.

This invention can further be represented in various ways, e.g., by the several combinatorial embodiments illustrated in Figs. 2A-6B. Each of the major elements introduced in Figs. 2A-6B are described in further detail in connection with Figs. 7-16. The antennas in each array can have arbitrary spatial placement, e.g., the array does not require a special antenna geometry in order to function correctly. Moreover, the antennas can be displaced in polarization as well as space.

Fig. 2A shows a point-to-point transmitter 30 comprising a stacked-carrier spread-spectrum (SCSS) transmitter bank 32 connected to a multi-element antenna array (AA) 34. A point-to-point receiver 36 comprises a multi-element antenna array (AA) 38 connected to a stacked-carrier spread-spectrum (SCSS) receiver bank 40. Each antenna array comprises a plurality of spatially separated antennas for transmitting and receiving data. Adaptive antenna array processing, e.g., adaptive linear combining and/or transmission over multiple spatially separated antennas, is not being combined here, or in Figs. 2B, 6A or 6B, with stacked carrier spreading and despreading. The array adaptation process is subsumed into the stacked carrier spreading and despreading operation.

Fig, 2B shows a network transmitter 42 comprising a stacked-carrier multiple-access (SCMA) transmitter bank 44 connected to a multi-element antenna array (AA) 46. A network

receiver bank 48 comprises a multi-element antenna array (AA) 50 connected to a stacked-carrier multiple-access (SCMA) receiver bank 52.

Fig. 3A shows a point-to-point transmitter 54 comprising s stacked-carrier spread-spectrum (SCSS) transmitter 56 connected to a time-division duplexer (TDD) 58. A point-to-point receiver 60 includes a time-division duplexer 62 connected to a stacked carrier spread-spectrum (SCSS) receiver 64.

Fig. 3B shows a network transmitter 66 which includes a stacked-carrier multiple-access (SCMA) transmitter 68 connected to a time-division duplexer (TDD) 70. A network receiver 72 comprises a time-division duplexer (TDD) 74 connected to a stacked-carrier multiple-access (SCMA) receiver 76.

Fig. 4A shows a point-to-point transmitter 78 that comprises a stacked-carrier spread-spectrum (SCSS) transmitter 80 connected to a code nuller 82. A point-to-point receiver 84 includes a code nuller 86 connected to a stacked-carrier spread-spectrum (SCSS) receiver 88.

Fig. 4B shows a network transmitter 90 that comprises a stacked-carrier multiple-access (SCMA) transmitter 92 connected to a code nuller 94. A network receiver 96 includes a code nuller 98 connected to a stacked-carrier multiple-access (SCMA) receiver 100.

Fig. 5A shows a point-to-point transmitter 102 that includes a stacked-carrier spread-spectrum (SCSS) transmitter 104 connected to a widely separated frequency channelizer 106. A point-to-point receiver 108 comprises a widely separated frequency channelizer 110 connected to a stacked-carrier spread-spectrum (SCSS) receiver 112.

Fig. 5B shows a network transmitter 114 that includes a stacked-carrier multiple-access (SCMA) transmitter 116 connected to a widely separated frequency channelizer 118. A network receiver 120 comprises a widely separated frequency channelizer 122 connected to a stacked-carrier multiple access (SCMA) receiver 124.

Fig. 6A shows a point-to-point transceiver system 126 wherein a stacked-carrier spread-spectrum (SCSS) transmitter bank 128 is connected to a synchronized time-division duplexer (TDD) bank 130 that is connected to both a multi-element antenna array (AA) 132 and a stacked-carrier spread-spectrum (SCSS) receiver bank 134 with a retro-adaptor 136 in control of the stacked-carrier spread-spectrum (SCSS) transmitter bank 128.

Fig. 6B shows a network system 138 that includes a stacked-carrier multiple-access (SCMA) transmitter 140 connected to a synchronized time-division duplexer (TDD) 142 that is connected to both a multi-element antenna array (AA) 144 and a stacked-carrier multiple-access (SCMA) receiver bank 146 with a retro-adaptor 148 in control of the stacked-carrier multiple-access (SCMA) transmitter bank 140.

Fig. 7A illustrates a stacked-carrier spread-spectrum (SCSS) transmitter 150 similar to those included in Figs. 2A, 3A, 4A, 5A, and 6A. The SCSS transmitter 150 includes a digital-to-analog converter (DAC) 152 that converts incoming digital data to an analog signal for transmission. Analog information for transmission may be input directly without the DAC 152. Two or more channels (e.g., 1,...,k) are included that each modulate corresponding radio frequency carriers in an up-conversion process. For example, each up-conversion channel comprises an inphase (I) mixer 154 and a quadrature (Q) mixer 156 connected to a 90° phase shifter 158 and a local oscillator (LO) 160. The modulating information therefore controls the amplitude of the inphase and quadrature phases of the AM carrier radio frequency. A pair of gain-controlled amplifiers 162 and 164 permit the independent adjustment of each of the inphase and quadrature amplitudes before recombining by a summer 166. A bandpass filter (BPF) 168 strips off out-of-band signals that could interfere with adjacent channels. A final summer 170 combines the signals from all of the channels and produces a transmitter output, e.g., that is then applied to an antenna. A spreading gain generator 172 periodically issues a parallel output that controls all the gain-controlled amplifiers 162 and 164 in every channel as a group. Each control signal to each gain-controlled amplifier 162 and 164 may comprise a signal digital line for one-bit on/off control, multi-bit parallel digital control for discrete gray-scale settings or an analog control for continuously variable gain settings.

An obvious variation on the analog circuitry shown in Figs. 7A and 7B for the transmitter 150 and receiver 180 is to use an all-digital transmultiplexer ("transmux") design, e.g., with discrete digital logic or with a digital signal processor.

A preferred alternative to the direct or transmux spreading and despreading approaches illustrated by example in Figs. 7A and 7B is the discrete multitone (DMT) method of orthogonal frequency division multiplexing (OFDM) described herein.

Referring to Fig. 7A, in operation of the transmitter 150, some spreading gain output from the spreading gain generator 172 is likely to exist which is more readily received by an intended receiver unit than that which would be obtained by using a different spreading gain output. The intervening radio communication environment between the transmitter and receiver will ordinarily attenuate or interfere with some phases and frequencies more than others. The radio communication environment contains co-channel interference, additive inter-network, intra-network and jamming/overlayered signals which are more readily circumvented with spreading code that cannot be eliminated at the receiver. The spreading gain output therefore has the ability to compensate for the effects of the intervening radio communication environment, both channel distortion and co-channel interference. The optimum spreading gain outputs that should

be generated at any one time can be fixed in patterned sequences according to time or place, or adjusted according to the results obtained from some sort of measurement related to the communication quality, e.g., reverse channel data. The spreading code provides for the compensation for co-channel interference sources, as well as channel distortion.

Fig. 7B illustrates a stacked-carrier spread spectrum (SCSS) receiver 180 similar to those included in Figs. 2A, 3A, 4A, 5A, and 6A, and complementary to the transmitter 150 shown in Fig. 7A. The SCSS receiver 180 accepts analog signals at a splitter 181 which drives parallel several frequency-separated channels. A typical channel comprises a bandpass filter 182, a splitter 183, an inphase gain-controlled amplifier 184, a quadrature phase gain-controlled amplifier 185, a pair of phase detectors 186 and 187 driven by a phase shifter 188 and a local oscillator 189 and an analog-to-digital converter (ADC) 190 that combines back all the receiver channels into a digital signal. Each down-conversion channel comprises inphase (I) mixer 186 and quadrature (Q) mixer 187 connected to the 90° phase shifter 188 and local oscillator (LO) 189. A despreading weight generator 191 is connected to control the individual inphase and quadrature amplifiers 184 and 185 of each channel.

A base station 230 is shown in Fig. 8. For "code-nulling," the despreading weights are adapted to maximize the signal-to-interference-and-noise ratio of the despread message sequence in the preferred embodiment; and to introduce directivity and retro-directivity, by noting that the spreading gains are derived from the locally adapted despreading weights in the preferred embodiment. The base station 230 is similar to base station 11 (Fig. 1) and comprises an antenna array 232 for directional radio communication with remote units by beam forming, a transmit/receive (T/R) front end 234, a frequency channel bank 236, a data cell mapper 238, a weight adaptation algorithm generator 240, a multi-antenna multi-link despreader 242, a delay and Doppler estimator 243, a delay and Doppler equalizer bank 244 and a symbol decoder bank 246, e.g. a Trellis decoder, that outputs several recovered baseband data channels. None, some or all of the antennas in antenna array 232 may be polarization diverse (e.g., antenna 233).

Several outgoing baseband data channels are connected to a symbol encoder bank 248, e.g. a Trellis encoder. From there, the transmission involves a delay and Doppler preemphasis bank 250, a multi-antenna multi-link spreader 252, an antenna and frequency channel mapper 254, a transmit/receive compensation bank 255 connected to a transmit/receive compensation algorithm generator 256, and an inverse frequency channelizer bank 257 connected to the T/R front end 234. A transmit/receive packet trigger 258 receives GPS time transfer information and controls the interleave and duration of individual transmit and receive times in the T/R front end 234. The base station can also have as few as one antenna element in its array. In a preferred

embodiment, the base station uses a packetized time-division duplex DMT or OFDM modulator and demodulator to perform the inverse frequency channelizer and frequency channelizer operations.

For more information on the alternative use of Trellis coded modulation, see, Boulle, et al., "An Overview of Trellis Coded Modulation Research in COST 231," IEEE PIMRC '94, pp. 105-109.

A remote unit 260 is shown in Fig. 9 in a preferred embodiment. The remote unit 260 is similar to remote units 12-17 (Fig. 1) and comprises an antenna array 262 for radio communication with the base station by combined spatial and spectral diversity, a transmit/receive (T/R) front end 264, a frequency channel bank 266, a data cell mapper 268, a weight adaptation algorithm generator 270, a multi-antenna despreader 272, a delay and Doppler estimator 273, a delay and Doppler equalizer 274 and a symbol decoder 276, e.g., a data decoder, that outputs a recovered baseband data channel. None, some or all of the antennas in antenna array 262 may be polarization diverse (e.g., antenna 263).

The outgoing baseband data channel is connected to a symbol encoder 278, e.g. a data encoder. The transmission further involves a delay and Doppler preemphasis unit 280, a multi-antenna spreader 282, an antenna and frequency channel mapper 284, a transmit/receive compensation bank 285 connects to a transmit/receive compensation algorithm generator 286, and an inverse frequency channelizer bank 287 connected to the T/R front end 264. A transmit/receive packet trigger 288 receives GPS time transfer information and controls the interleave and duration of individual transmit and receive times in the T/R front end 264.

The remote unit can have as few as one antenna element in its array. The number of antennas at each remote unit can vary from unit to unit. This can allow the remote units to have a variable cost, based on the importance or data rate employed by a given unit. The remote units can use different spreading rates. They can spread their data over different subsets of the frequency channels used by the base station transceiver. In a preferred embodiment, the remote unit uses a packetized time-division duplex DMT or OFDM modulator and demodulator to do the inverse frequency channelizer and frequency channelizer operations. A difference between the base station and the remote unit is that the base station transceiver signals from multiple nodes, e.g, multiple access. Each remote unit transceives only the single data stream intended for it. Channel equalization techniques and code nulling are limited methods for adapting the spreading and despreading weights.

Fig. 10 illustrates a multi-antenna transmit/receive module 290. The module 290 includes a multi-element antenna array 291 with each element connected to a corresponding single-

channel T/R module 292, e.g., four in number. Each T/R module 292 is connected to a packet trigger 293, a receiver calibration generator 294, a local oscillator 295 and a system clock 296. These, in turn, are driven by GPS clock and Doppler correction signals. Each T/R module 292 comprises a T/R switch 297, an intermediate frequency (IF) down-converter 298, an analog-to-digital converter (ADC) 299, a digital-to-analog converter (DAC) 300, an IF up-converter 301 and a power amplifier (PA) 302. Receive weight information is learned during the receive process and is used in the transmit process to set the relative transmit powers applied to each antenna element, e.g., to compensate for channel fade or interference. It should be noted that the transmit/receive module must separately excite both polarizations if the base station is polarization diverse.

The receive and transmit time slots are triggered at particular times, which may be pseudo-randomly determined, according to an independent source of accurate, universally accessible time, e.g., from the global positioning system (GPS) maintained by the United States Department of Defense. Such GPS times are derived from a navigation system residing on board a communication platform such that the receiver side of each T/R module 292 always knows what time slot a packet corresponds to. The GPS time is also used to derive the local oscillator and ADC/DAC clocks used in the system. The receiver side is not necessarily synchronized to the remote transmitting source. In particular, the range, propagation delay and Doppler shift between the communicators need not be known by the receiver system prior to the reception of a first data packet. However, the range, velocity, delay and Doppler shift between the communicators may be known to some degree of accuracy in certain applications. The range, propagation delay, and Doppler shift between the communicators does not need to be known prior to reception of the first data packet.

The calibration mode is optional, and is only used on an as-needed basis. For example, intermittently, at the beginning of a given transmission, or when internal diagnostics indicate that such calibration is required.

The encoding, spreading, and modulation operations, e.g., as shown in Fig. 11, are preferably mirrored by analogous demodulation, despreading, and decoding operations, e.g., as in Fig. 12. The dataflow in Fig. 11 is reflected in Fig. 12 as a signal flow, e.g., the same data flows are in both Figs. 11 and 12, with the adders in one Figure replaced by fan outs in the other Figure. Such symmetry is exemplified by DMT modulators and demodulators, frequency mapping and inverse mapping operations, spreading and despreading operations, and code-gated spreading and despreading operations. The structure of the spreader mirrors the structure of the

despreader. Prior art CDMA transceivers do not have such symmetry. Thus, such symmetry is a critical feature in embodiments of this invention.

Fig. 11 illustrates a discrete multitone stacked-carrier-spread-spectrum (SCSS) modulator used for frequency channelization in the embodiment 300. A frame generation command from navigation and coding system 302 causes a signal modulator 304 to encode ephemeris, position, velocity, acceleration, and other messages into a K_{cell} symbol data vector. These symbols are then used to modulate a set of baseband tones or fast Fourier transform (FFT) bins. In a spreader 306, the K_{cell} baseband tones are replicated over K_{spread} separate spreading cells, multiplied by a separate spreading gain for that antenna "1" and frequency cell "h" complex, e.g., complex constant equally multiplying each symbol in the cell, and passed to a time-multiplexer that combines the cells into a K_{active}-long vector of complex data, where $K_{active} \geq K_{cell} * K_{spread}$. This complex data vector is passed to a zero-padded inverse-FFT operator 308 that converts the data vector directly to $K_{FFT} \geq (1+SF) \cdot K_{active}$ real-IF time samples, wherein "SF" denotes the "shape factor," or ratio of stopband to passband for this system. The first $E_{roll} \bullet K_{FFT}$ samples at this time sequence are then repeated 310 to form a $K_{packet} = (1+ E_{roll}) \cdot K_{FFT}$ - long data sequence. A multiplier 312 multiplies this by K_{packet}-long data from a Kaiser-Bessel window 314 to generate the final sampled signal. The sampled signal is then passed to a digital-to-analog converter resulting in a $T_{packet} \cdot K_{packet}/f_s$ -long data burst, passed to frequency up converter and to the communication channel where f_s is the complex sample rate of the DPC/DNC nodules. The parameters used to reduce the features of the transmitted signal are all coordinated to GPS time, such that the nodes in the communication network transmit simultaneously. This process its repeated for each antenna in the system.

A symbol encoding at the baseband tones is included in the baseline system 300. Each K_{cell} data bit modulates a separate tone in the signal baseband, such that a tone is phase modulated by O or 180° if the data bit modulating that tone is equal to zero or one, respectively. Such tone modulation is highly efficient in terms of allowable transmit power. It offers a vulnerability to radiometric detection techniques, and it allows reliable demodulation of the transmitted bit sequences at E_b/N_0 of as low as three dB. The BPSK format allows for the use of powerful and sophisticated methods to remove timing and carrier offsets from the despread signal, when based on the conjugate self-coherence of the tone phase sequence.

Such operations are for a single antenna, e.g., using a different complex spreading gain g_{k1}, for each frequency cell k and antenna 1 employed by the transceiver. The passage uses the packet extension factor e_{roll} and the packet sample length $K_{packet} = (l+e_{roll})K_{FFT}$ samples prior to the digital-to-analog conversion operation ($T_{packet} = (l+e_{roll})T_{FFT}$ time duration after the DAC

operation). The spreading gains g_{kl} can be determined via number of means, for example, via codebook, randomly, pseudo-randomly, or adaptively, based on the despreading weights w_{kl}.

The number of information bits per data symbol is K_{bit}. The BPSK is a simple encoding strategy where encoding is ignored and $K_{bit} = 1$. The platform ephemeris, position, velocity, and acceleration information, are examples of data that could be transmitted in some applications. BPSK is a preferred modulation for applications where the data rate is not the primary concern of the system.

The delay and Doppler preemphasis operation is optionally included in alternative embodiments. Such can be included after the initial packet to remove the effects of delay and Doppler shift at the intended receivers of the signal being transmitted from the DMT modulator. This operation can simplify the design of the transceivers in the network, e.g., by allowing the delay and Doppler removal operations to be concentrated at the base stations in the network.

As a generalization of the spreading concept to multiple access transceivers, a separate set of spreading gains ($g_{kl}(m)$) can be used to spread the data symbols intended for user m in the multi-user transceiver.

Fig. 12 shows an all-digital fully-adaptive despread and beam-forming receiver 320. For a background of this technology see, Tsoulos, et al., "Application of Adaptive Antenna Technology to Third Generation Mixed Cell Radio Architectures," March 1994, IEEE #1-7803-1927, pp. 615-619. A frame reception command from a receiver navigation and coding system 322 causes a signal demodulator 324 to collect and convert from analog to digital a series of T_{gate}-long transmit frames from a K_{array} of array antennas 326, where T_{gate} is the time duration spanned by K_{gate} samples. This includes a T_{guard}-long guard time slot to account for the unknown propagation delay between the transmit and receive links ($T_{gate} = T_{packet} + T_{guard}$), where T_{packet} is the time span of the packet and T_{guard} is the time interval spanned by K_{guard} samples. A K_{gate}-long digitized data frame is output from each ADC and is then passed to a windowed, zero-padded sparse FFT 328 that converts the packets to the frequency domain with each tone separated by an integer number of FFT bins.

The FFT bins are passed to a demultiplexer 330 which removes any unused FFT bins from the received data set, and groups the remaining bins into $K_{cell}x(K_{spread} \cdot K_{array})$ data matrices containing the tones received over each transmitted spreading cell, where K_{spread} is the frequency spread factor, K_{cell} is the number of symbols per pre-spread data cell, and K_{array} is the number of antennas. Each of the spread data cells are passed through a bank of linear combiners 332 that remove the co-channel interference covering each cell and despreads the original baseband symbol tones from the received data set. The combiner weights are adapted using a code-gated

self-coherence restoral method that simultaneously despreads the received data signal and does frequency-dependent multi-antenna reception and spatial filtering of the spread signal-of-interest.

The combiner weights are used to construct a set of transmit weights to be used in any subsequent return transmission. Such tones are then passed to a delay and Doppler equalization unit 334 that estimates and removes the Doppler shift (non-integer FFT bin-shift) and message propagation delay (phase precession) from the received data set. A symbol demodulator 336 estimates the transmitted message symbols.

As a consequence, the received data packet transmitted from each user is despread and extracted from the received interference environment. The processor does not require fine timing/carrier sync to the transmitter until after the baseband signal has been despread with a high signal-to-interference-and-noise ratio, even in the presence of strong noise and co-channel interference.

The K_{cell} symbols transmitted from user m are extracted from the channel at the receiver by weighting each of the K_{cell} tones received on frequency cell k and antenna 1 by the same complex despreading weights $w_{k1}(m)$, and then adding the cells together on a tone-by-tone basis, such that tone q in each received frequency cell is summed over all of the $K_{spread} \cdot K_{array}$ frequency cells and antennas used by the system.

Each multi-element transceiver preferably has the minimum number of combined spatial and spectral degrees of freedom $K_{array} \cdot K_{spread}$ to successfully null any non-stacked carrier interference sources invading on each frequency cell. Any excess remaining degrees of freedom are used to improve the SINR of the despread baseband signal or to separate overlapping stacked carrier signals. The multi-element despreader weights are then adjusted to maximize the power of the despread baseband signals. This leads to code nulling solutions that can be considerably more powerful than conventional methods of despreading. The ideal despreader adjusts the despread weights to null non-stacked-carrier interference sources down to a noise floor over each frequency cell, and simultaneously enhances the SINR of the despread signals. The multi-element despreader also preferably directs significantly weaker nulls against interference sources with weaker radio signals in a given frequency cell. Soft nulls can therefore be directed at interference sources received with weaker power in a given frequency cell. For example, a weaker null can be directed at the far edge of an interference source passband, if the interference source spectrum has a particularly weak value at those frequencies.

In general, the despreader weights including adaptive antenna arrays significantly improves the quality and capability of the signal transmission and reception operation. For the receiver

side of a system, blind or non-calibrated methods can be used to direct near-optimal beams at signals-of-interest, and to simultaneously direct nulls at jamming signals.

In general, the despreading weights are adjusted to maximize the signal-to-interference-and-noise ratio (SINR) of the despread baseband signals, e.g., estimated data symbols. This typically results in a set of code-nulling despreading weights that are significantly different than the spreading gains used to spread the baseband signals at the other ends of the link. In particular, such resultant despreading weights will simultaneously remove channel distortion, such as selective gains and fades caused by multipath. Despreading provides an optimum tradeoff between nulling interferers received by the transceiver, by maximizing signal-to-interference ratio, and maximizing the signal-to-noise ratio (SNR) of the despreader. The despreading codes in conventional DSSS and CDMA communication systems are set equal to the spreading codes at the other ends of the link and only maximize the SNR of the despread baseband signals.

Such operation is performed blindly in the preferred embodiments of the invention, the transmit spreading gains and channel distortions are not known at the despreader. This simplifies the protocols used within the network by allowing use of unknown spreading gains at the transceivers in the network. This also allows the use of adaptively determined spreading gains that are continually optimized to mitigate the noise, interference, and channel distortion encountered by the transceiver over the course of a transmission.

Such approach provides an upgrade to multi-element SCMA or SCSS transceivers using antenna arrays, by not requiring any qualitative change in the spreading, despreading, or gain/weight adaptation algorithms. A difference lies in the multi-element transceiver in the dimension of the multi-element spreading and despreading operations. However, the multi-element transceiver has greater capacity due to the larger degrees of freedom that it can use to separate SCSS signals. The range and/or immunity to intercept by radiometric detection means is increased due to its ability to steer spatial beams at other communicators in the network. The immunity to jamming from non-SCSS signals is also improved due to the ability to spatially null such signals, even if they are arriving from over broad frequency ranges.

Rapid-convergence methods that function over single data packets can also be combined with frequency-channelized signals-of-interest or processor structures to allow frequency-selective nulling of the interference source signals without the need for array calibration data or the need to know or estimate the direction-of-arrival of the signal-of-interest or interference source signals. The system 10 (Fig. 1) can therefore detect and demodulate data packets in highly dynamic environments where the channel geometry is changing significantly between packets. As such, a processor could operate in the typical overloaded environments where the

number of interferers are not less than the number of antennas in the antenna array for the receiver.

On the transmitter side of a system, directive or retrodirective adaptation methods can be used to either direct returning signals-of-interest back to the transmit source with maximum power and/or minimum transmit radio signals (directive mode), or to jointly direct the returning signals-of-interest back to the transmit source with minimum radiation in the direction of the interfering sources (retrodirective mode).

The directive mode is useful in applications where compatibility with non-SCSS interferers is not a primary concern to the communicator, or where interference transmission and reception platforms are not likely to be placed in the same location. This mode is also useful in applications where the communication platform is subject to heavy non-SCSS interference, such that maximum power must be delivered to the other end of the communication link.

Processors can be used to accurately measure the received signal-of-interest steering vector and direct a maximal beam back to the other end of the communication link without knowledge of the received signal-of-interest direction-of-arrival, even though the interference sources completely cover the signal-of-interest passband and packet interval. The system 10 (Fig. 1) preferably delivers a factor-of-K_{array} more power to the other end of the communication link, providing the system with additional immunity to any jamming. This can be accomplished even if the other end of the communication link is transmitting and receiving over a single antenna. Conversely, the system 10 (Fig. 1) can maintain the communication link using a factor-of-K_{array} less power. This reduces the geographical area within which the system can be detected by adversaries by a factor of K_{array}.

The retrodirective mode, embodied in Figs. 6A and 6B as retro-adapters 136 and 148, is useful in applications where interceptors are placed in the same location with the interference sources, e.g., in order to assess the effectiveness of the jamming strategy. This strategy is most useful in underloaded environments where broadband nulls can be directed at the interference sources.

Fig. 13 illustrates a single-frame digital multitone (DMT) modulation and spreading format 340. The format 340 is used for an example environment with $K_{cell} = 6$ and $K_{spread} = 4$, and with each spreading cell separated by two FFT bins, so $K_{space} = 2$. The six data bits to be transmitted are first transformed to a set of ± 1 data symbols. The symbols energize six baseband FFT bins that are replicated over four cells of FFT bins, e.g., spreading cells, with a separate complex weighting g_k on each cell. The complex weights are the spreading gains, and are randomly or pseudo-randomly set over each data packet. The spreading is performed in the

frequency domain by multiplying the time-domain representation of the baseband signal by a set of superimposed, or stacked, complex sinusoid carrier waves. In practice, the spreading is done by simply energizing the bins of a large FFT, at a considerable savings in computational complexity for moderate output FFT sizes. A Kaiser-Bessel window with ß = 9 is used in this invention to "fill out" the space between the tones without subjecting those tones to unacceptable interference from adjacent tones, e.g., intertone interference. In particular, the high value of ß provides acceptable interference between adjacent tones and extremely low interference between farther-out tones.

Non-blind or calibrated techniques use a knowledge of the baseband data sequence or channel distortion and spreading gains to develop ideal weights based on optimum signal estimation methods; for example, least-squares techniques. Blind or non-calibrated techniques use more general properties of the baseband data signals to adapt the despreading weights. Mixtures of these techniques that use known and unknown components of the baseband signal and/or the transmission channel can also by used to construct an effective solution. Examples of blind techniques that are particularly useful include constant-modulus, multiple-modulus, and decision-direction techniques. Such use properties of the message symbol constellation to adapt the despreading weights. A number of methods can be used to adapt multi-element despreader weights in the demodulator 332 (Fig. 12). First, there are dominant-mode prediction (DMP) methods which take advantage of the known packet arrival times or known spreading parameters of the discrete multitone stacked carrier signal. Second, there are code-gated self-coherence restoral (SCORE) methods, which take advantage of the known self-coherence, or non-zero correlation between the spectrally separated signal components, in the discrete multitone stacked carrier signal.

Of these two basic kinds, the self-coherence restoral technique has the greatest utility for single-packet acquisition and detection of the discrete multitone stacked carrier signals.

Conventional spectral and other types of self-coherence restoral take advantage of the known spectral and/or conjugate self-coherence property. This is a non-zero correlation between frequency-shifted and/or conjugated components of a given communication signal. Blind methods do not require any prior knowledge of the content of the signals-of-interest or their directions-of-arrival. So no specific receiver calibration information is needed to train the antenna array for the receiver. Instead, the blind method uses its own local knowledge of the specific frequency-shifts that the signals-of-interest are correlated over. See, B. Agee, S. Schell, W. Gardner, "Self-Coherence Restoral: A New Approach to Blind Adaptation of Antenna Arrays," in Proceedings of the Twenty-First Asilomar Conference on Signals, Systems and

Computers, 1987. And see, B. Agee, S. Schell, W. Gardner, "Self-Coherence Restoral: A New Approach to Blind Adaptive Signal Extraction Using Antenna Arrays," IEEE Proceedings, Vol. 78, No. 4, pp. 753-767, April 1990. See also, B. Agee, "The Property Restoral Approach to Blind Adaptive Signal Extraction," Ph.D. Dissertation, University of California, Davis, CA, 1989.

In a double side band amplitude modulated signal, the real-IF representation of any such signal possesses conjugate symmetry about both its carrier frequency, due to the double side band amplitude modulated modulation format, and DC, due to the real-IF representation. These symmetries offset each other, causing the negative and positive frequency components of the signal to be equal to each other. This perfect spectral self-coherence is observed by computing the correlation coefficient between the double side band amplitude modulated signal-of-interest and a replica of itself that is frequency-shifted by twice the carrier. The frequency-shift operator mixes the negative frequency components to the frequency band occupied by the positive frequency components, causing the correlation coefficient to possess a non-zero value. Such non-zero value occurs only when this value of frequency shift is applied to the replica. The correlation coefficient is less than unity in this example. A unity correlation coefficient is obtained by filtering out the extraneous non-overlapping signal-of-interest radio signals in the original and frequency-shifted double side band amplitude modulated signals.

In Fig. 14, a cross-self-coherence restoral (SCORE) processor 350 is used to do a restoration that is applied to a multi-antenna received data signal x(t). The processor 350 first passes the received data through a series of filtering, frequency-shifting, and, optional, conjugation operators, resulting in a signal u(t) that is only correlated with the signals targeted by the processor. The original and processed signals x(t) = u(t) are then passed through a pair of beam-and-null steerers (linear element combiners) 352 and 354 that are adapted to maximize the correlation coefficient between the combiner output signals y(t) = w"x(t) and r(t) = c"u(t). The control parameters used to target the processor are the filter operator, typically set to a delay operator, frequency-shift value α, and conjugation flag (*). The processor parameters are set to values that yield strong correlation coefficients in the absence of interference, e.g., in the transmitted signal of interest to the processor.

Figs. 15 and 16 show the code gating operation used in the general code-gated SCORE operation. Certain code-gating configurations require some significant modification to the spreader and despreader data flow, and therefore structure. These illustrate one method for enabling the code-gated SCORE despreader adaptation algorithm described herein. Other methods exist that apply the code-gating across packets or internally with frequency cells, rather

than across frequency cells. For example, by repeating the data symbols with a gating code applied across the K_{cell} baseband symbols over the even packets, and consequently do not affect the data flow through the spreader and despreader.

• Partition the spreading cells into $K_{part} \geq 2$ subsets, K_{SCORE} cells/subset —$K_{part} = K_{spread}$, cells processed in individual subsets, $K_{SCORE} = 1$ cells/subset

—$K_{part} = 2$, cells separated, into <u>even</u> and <u>odd</u> subsets, $K_{SCORE} = K_{spread}/2$ cells/subset —$K_{part} \cdot K_{SCORE} = K_{spread}$ in each case •Use same code key for every cell in subset: —$c(n;K_{part}l+k) = c(n;k)$, $k=O, \ldots, K_{part}-1$, $l=0, \ldots, K_{SCORE}-1$ —Alternate form: $c(n;k) = c(n;(k)K_{part})$, $k = 0, \ldots, K_{spread}-1$, $(k)K_{part}$ =modulo-K_{part} of k Transmit over multiple SCSS subchannels (stacked-carrier multiple access), using different code keys with <u>same structure</u> on each subchannel —Allows separation of $K_{array} \cdot K_{SCORE}$ SCSS subchannels per subset (code-nulling performed) —Enables <u>higher data rates</u> per user (multiple subchannels per user) —Enables <u>multiple access communications</u> (multiple users communicating with cell) —Enables rejection of $K_{SCORE}-1$ SCSS interferers (cellular communications) —Requires higher time bandwidth product to achieve same misadjustment level •In practice, K_{part} adjusted to specific application $K_{part} = K_{spread}$ in asynchronous point-to-point links, cellular overlay systems where non-SCSS interference high, fast convergence time important $K_{part} = 2$ in point-to-multipoint links, SCMA systems where SCSS interference high

Code-gated self-coherence restoral takes advantage of self-coherence information that has been deliberately added by the communication system to facilitate the adaptive spreader, but that cannot be discerned without access to the gating information in the communication network. Two code-gated SCORE methods are included in this invention.

A preferred self-coherence restoral method for multiple access communications includes applying unique code gating operations to the baseband message signals prior to the spreading operation, which is uniquely determined for each link in the system. For example, where the frequency cells are broken up into two subsets of cells, even and odd, with the code key applied over only the odd cells, as represented in Figs. 15 and 16. The data symbols are spread over the even cells using the method shown in Fig. 11, and related text.

A similar spreading operation can be applied over the odd cells. However, the data symbols transmitted over these cells are first subjected to a code gating operation where they are multiplied by a constant modulus code key $c(m) = [c_4(m)]$ that is different for each user m in the network. This operation is reversed at the multiple access despreader. The odd frequency cells are multiplied by the conjugate of the code key $c^*(m)$ after the despreading operation, but before combining the outputs of the despreaders employed on the even and odd frequency cells. At the single-user (SCSS) transceiver, the code gating operation is only performed for the single code key employed by the SCSS transceiver. During single-packet acquisition operations, the despread (conjugated) code key is applied to each of the odd received frequency cells and on each of the transceiver antennas, that is, before the linear combining operation.

The effect of the code gating operation is to cause the signal transmitted with that code key to have unity correlation coefficient between the even and odd frequency cells after the odd frequency cells are multiplied by the despread code key. Conversely, the same code gating operation will cause all of the other signals transmitted using different code keys to have low correlation coefficients between the even and odd frequency cells. This condition will hold regardless of the (assumed unknown) delay and Doppler shift inflicted on the received signals. The resultant signal can then be input directly to the cross-SCORE algorithm shown in Fig. 14, where x(t) is replaced by the even (not gated) frequency cells and u(t) is replaced by the odd (gated) frequency cells, and where t refers to the symbol index $q = 1, \ldots, K_{cell}$ rather than a time index. The despreading weights are adapted to maximize the correlation coefficient between the outputs of the despread linear combiners applied to the even and odd frequency cells.

Such method provides for unambiguous detection and despreading of any link in the network, based only on the known code key for that link. At the single-user SCSS transceivers, the transceiver despreads only the link that it is communicating over, without the need for additional operations to acquire the link and verify that it is carrying the correct signal. The link is automatically reacquired if it is temporarily lost due to adverse channel conditions, e.g., "port shuffling" that occurs over long transmissions. At the multi-user SCMA transceivers, the

method allows unambiguous detection, despreading, and identification of every link supported by the transceiver, without port swapping or shuffling as the channel conditions vary, based only on the known code keys used by the nodes linked to that transceiver. The code keys provide some privacy by the scrambling included in the code-gating operation.

The basic code-gated SCORE method can also be generalized in many ways. In particular, code keys may be applied to the even as well as odd frequencies, in order to provide enhanced security and decorrelation between the frequency cells. The code gating may also be applied over time rather than frequency, by transmitting data symbols on consecutive packets with code gating omitted during the even packets and performed over all of the frequency cells during the odd packets. If the spreading codes are held constant over these pairs of packets, this approach allows the use of the more powerful auto-SCORE method to adapt the despreading weights.

•More powerful algorithm allowed in some environments
—Channel response approximated as identical or nearly identical (differs by complex scalar) on each spreading subset
—Background interference approximated as identical on each spreading subset
•Leads to maximum likelihood estimator
—Spreading gains forced identical or nearly identical on each spreading subset
—Despreading weights forced identical or nearly-identical on each spreading subset
—Despreading weights set to dominant mode(s) of auto-SCORE eigenequation
•Has advantages over cross-SCORE eigenequation
—Lower complexity
—Lower misadjustment at same time-bandwidth product
—Forces max-SINR equal on each subset in network applications – no asymptotic misadjustment can occur
•Some disadvantages
Sensitive to modeling error if channel response truly unequal on each subset
Requires tracking/removal of timing and/or Doppler during despreading operation (algorithm typically fairly simple)

Larger numbers of frequency or packet subsets may also be used in the system, with a separate set of code keys employed over each subset. In this case, the despreader uses a generalization of the cross-SCORE method that derives from a super-vector interpretation of the cross-SCORE eigenequation. See, B. Agee, "The Property-Restoral Approach to Blind Adaptive

Signal Extraction," in Proc. 1989 CSI-ARO Workshop on Advanced Topics in Communications, May 1989, Ruidoso, NM; and, B. Agee, "The Property Restoral Approach to Blind Adaptive Signal Extraction," Ph.D. Dissertation, University of California, Davis, CA, June 1989. As the number of frequency subsets rises, the number of multiple access communications that can be supported by the transceiver drops, but the stability of the weight computation improves, and the noise reduction and nonstacked-carrier nulling capability of the algorithm remains unchanged. In the limit as the number of frequency subsets equals the spreading factor K_{spread}.

The code-gated-self-coherence restoral method extracts the signal-of-interest baseband directly from the channelized data super-vector, using the dominant mode of a multicell self-coherence restoral eigenequation. The method simultaneously does frequency-dependent spatial filtering, combines antenna elements within each cell on the spread signal-of-interest, and despreads the resultant data signal to combine the frequency cells.

The code-gated-self-coherence restoral method can operate effectively at positive or negative receive SINRs, as long as the maximum attainable despread and beam-formed SINR of the received data packet is positive. The method adapts the antenna array as an intrinsic component of the despreading, linear combining, operator. The same method is used for arbitrary numbers of antennas, including single antenna systems where K_{array} = one. The code-gated-self-coherence restoral method does not require a prior knowledge of the spreading gains or underlying message sequence at any point in its implementation. The method does not require a search over time or Doppler shifted frequency to despread a message sequence.

The dominant eigenvalue of the code-gated-self-coherence restoral eigenequation provides for detecting new signal packets when the communication link is first opened. The receiver functions in an "on-demand" basis, returning pulses back to the other end when a packet is transmitted in the communication channel.

Additional methods enhance or confirm detection of a discrete multitone stacked carrier data packet after the code-gated-self-coherence restoration. In particular, the detection reliability can be greatly enhanced by using the lesser eigenvalues of the code-gated-self-coherence restoral eigenequation to predict the mean and standard deviation of the maximum code-gated-self-coherence restoral eigenvalue. The true maximum eigenvalue is then decremented by the predicted mean and scaled by the predicted standard deviation, resulting in a much stronger trend corrected detection statistic.

Other methods use downstream despreading and demodulation operators to confirm packets detection during the code-gated-self-coherence restoration.

Initial Doppler recovery during acquisition of the first data packet uses a frequency-domain analog of a fractional-spaced equalizer by extracting the first data packet at the full remode of the receive-site FFT, and sub-sampling the resultant output signal down to the transmit-site frequency remode using a linear interpolation method. The linear combining weights re-sample the data to tone centers using some appropriate adaptation method. A least-squares property restoral algorithm, such as a constant modulus method, minimizes the variations in the modulus of the despread data symbols. The least-squares constant modulus method takes advantage of the property that the transmitted data tones have a constant modulus if they are generated using a BPSK modulation format, but this property is destroyed if the transmitted signal is subjected to a Doppler shift that is a non-integer multiple of the tone spacing. The least-squares constant modulus method restores this property to the despreader output signal. The overall technique operates in the presence of significant Doppler shifts and path delays. See, B. Agee, "The Least-Squares CMA: A New Approach to Rapid Correction of Constant Modulus Signals," in Proc. 1986, International Conference on Acoustics, Speech and Signal Processing, Vol. 2, pg. 19.2.1, April 1986, Tokyo, Japan.

Two general methods are useful in generating the antenna array weights for data transmission. Retrodirective transmission sets the transmit weights proportional to the conjugated receive weights, and the directive mode sets the transmit weights proportional to the conjugated packet steering vectors. The retrodirective mode is well suited to commercial telecommunications and military intraflight communication applications where the interfering signals may be the other members in a multipoint communication network.

The directive mode is most useful in applications where the covert nature is of primary concern to the communicator and the jamming and interception platforms are not likely to be placed in the same location. This mode is useful in applications where the communication platform is subject to heavy interferences, such that maximum power must be delivered to the other end of the communication link in order to communicate in the presence of the interfering radio emissions. However, this method does not possess the attractive property of directing energy away from other interferers in cooperative communication networks.

The directive mode also constitutes a multiplier adaption strategy that can greatly simplify the complexity of the despreader if large spectral spreading factors are employed by the airlink.

The retrodirective transmission mode is illustrated herein. The retrodirective mode sets the transmitter antenna array weights equal to the conjugated array weights computed during signal reception. If the transmit and receive operators are done over the same frequency band and any internal differences between the transmit and receive paths are equalized, then the transmitter

antenna array will have the same gain pattern as the antenna array for the receiver. The transmitter antenna array will estimate the direction of nulls in the direction of any interferers that were present during signal reception. The null depths to use for each are determined by the relative strength of the received interferers.

Throughout here, g_k is a K_{array} x 1 dimensional vector, and represents the multi-element spreading vectors employed at transmit over frequency cell "k". The multi-element despreading vectors used at the receiver over frequency cell "k" is represented by w_k is also a K_{array} x 1 dimensional vector.

Embodiments of the present invention are preferably configured to provide frequency selective transmit weights by spreading the transmit packet using a different set of K_{array} x 1 spreading (g_k) weights over each spreading cell. This sets the frequency-selective retrodirective transmit weights, by setting the (multi-element) spreading gain g_k proportional to the K_{array} x 1 linear combiner despreading weights w_k employed over each frequency cell during signal reception such that $g_k = \lambda w_k^*$. This mode is especially effective in environments dominated by broadband interference sources, since the resultant null depths will be limited by the antenna array dispersion realized over each frequency cell. In this case, the processor will null the interference source over frequency as well as space. The transmitter antenna array only nulls each interference source over the frequency cells occupied by that inference source. This is fine for receiving a signal-of-interest packet, but it is ineffective for transmitting a packet if the goal is to direct packet radio signals away from interference source locations over the entire packet passband. Such a goal cannot be met by any means if the number of partial-band interference sources equals or exceeds the number of elements in the antenna array.

The directive transmission mode sets the transmitter antenna array weights equal to the (conjugated) K_{array} x 1 packet steering vector. If the transmit and receive operators are done over the same frequency band, with appropriate equalization of any differences between the transmit and receive paths past the transmit/receive switch, then the resultant antenna array will direct maximum radio energy to the other end of the communication link or close the link with minimum transmit radio energy. The directive array typically ignores the locations of the interference sources, e.g., it implicitly assumes that the interceptors are anywhere in the field of view of the communication link.

The directive method can be implemented on a frequency-selective basis in the present invention. This can provide some benefit in exceptionally wideband communication links, e.g., due to large values of K_{spread} or over highly dispersive communication channels where the packet steering vector changes significantly over the packet passband. However, this is unimportant

because the maximum power mode is not strongly degraded by minor errors in the packet steering vector.

Such estimation error can be large if the communication link is heavily jammed, or if the packet steering vector must be estimated over short communication intervals, e.g., single packets. In particular, an overly simple method can cause the directive transmitter antenna array to point strong beams of transmission energy at the interference sources in the environment. A directive transmission method or packet steering vector estimator needs to be simple enough to implement inexpensively, but sophisticated enough to reliably operate under expected ranges of jamming and transmission scenarios.

Three general steering vector estimation methods are preferred. First, correlative methods, which estimate the packet steering vector using the correlation between the received and estimated packet data. Second, multicell ML-like methods, which estimate the packet steering vector using the maximum-likelihood (ML) estimator obtaining under appropriate simplifying conditions and in the presence of frequency-channelized (multicell) data. And third, parametric methods, which further refine the multicell estimators by constraining the packet steering vector using appropriate parametric models.

The correlative method is the simplest method of the three to estimate the packet steering vector. A weakness of this method is seen by considering the estimate obtained in the presence of a single interference source, the estimate reduces to the packet steering vector plus the interference source steering vector scaled by the cross-correlation between the received interference source and packet signals. The time-bandwidth product (samples) required to reduce this cross-correlation to zero is much larger than the 1/S of the interference source signal, for example, 1,000,000 samples if the interference source is fifty dB stronger than the packet signal. As a consequence, this method is usually inadequate.

The other two methods can overcome this limitation by using optimal maximum likelihood (ML) estimation procedures to estimate the packet steering vector. The resultant estimators can provide accurate steering vector estimators in the presence of broadband or partial-band interference sources, using simple (non-parametric) or sophisticated (parametric) steering vector models. Moreover, the performance of these estimators can be predicted using conventional Cramer-Rao bounds analysis.

A useful performance bound is derived for any non-parametric steering vector estimate obtained in a multicell environment. The received data is partitioned into K_{spread} separate frequency cells, each containing a known (or estimated) packet baseband scaled by an unknown complex steering vector and corrupted by additive complex Gaussian interference. The steering

vector in the Pe cell is modeled by $a_k = g_k$ a where "a" is the (frequency-independent) packet steering vector and g_k is the scalar received single-antenna packet spreading gain obtaining over the k^{th} spreading cell. The complex Gaussian interference is assumed to be independent from cell to cell, and temporally-white with mean zero and unknown autocorrelation matrix $\{\mathbf{R}_{i_k i_k}\}$ in the k^{th} cell.

The packet steering vector **a** is assumed to be an arbitrary complex of dimension K_{array} vector of dimension K_{array}, e.g., **a** is not constrained to adhere to any parameterized model set (for example, an array manifold parameterized with respect to azimuth and elevation). Steering vector estimates developed using this model are e.g., non-parametric techniques. See, H. Van Trees, Detection, Estimation, and Modulation Theory, Part I, New York: Wiley, 1968. Using Cramer-Rao bounding theory, any unbiased estimator of **a** will have an estimation accuracy (mean-squared error) bounded by the Cramer-Rao bound given. The matrix **R** is interpreted as a generalized "average" of the interference autocorrelation matrices $\{R_{i_k i_k}\}$ equal to the inverse of the averaged inverse autocorrelation matrices.

In the preferred embodiment, the spatial steering vector a and the spectral spreading gains (g_k) are iteratively computed using the formula

$$R = \left(\sum_{k=1}^{K_{spread}} |g_k|^2 R_{H_k H_k}^{-1} \right)^{-1}$$

$$a = R \left(\sum_{k=1}^{K_{spread}} g_k w_k \right)$$

and

$$g_k = \frac{w_k^H a}{a^H R_{H_k H_k}^{-1} a}$$

where $R_{H_kH_k}$ is the data autocorrelation matrix measured over an adaptation block at spectral cell k, and w_k are the spatial component of the despreading weights employed at spectral cell k. The steering vector and spreading gains can also be used to compute improved despreading weights w_k, which can then be used in a multiple despreading procedure that performs spatial processing (linear combining at each frequency cell) followed by spectral processing (linear combining over frequency cells).

The stacked-carrier spread spectrum radio communication devices shown in Figs. 1-14 in combination with code nulling techniques represent alternative embodiments of the invention. Code nulling interference cancellation techniques can be effectively combined with stacked-carrier spread spectrum techniques. For more on code nulling design, see, Brian Agee "Solving the Near-Far Problem: Exploitation of Spatial and Spectral Diversity in Wireless Personal Communication Networks," *Wireless Personal Communications*, edited by Theodore S. Rappaport, et al., Kluwer Academic Publishers, 1994, Ch. 7. And see, Sourour, et al., "Two Stage Co-channel Interference Cancellation in Orthogonal Multi-Carrier CDMA in a Frequency Selective Fading Channel," IEEE PIMRC '94, pp. 189-193. See further, Kondo, et al., "Multicarrier CDMA System with Co-channel Interference Cancellation," March 1994, IEEE, #0-7803-1927, pp. 1640-1644.

The basic stacked-carrier spread spectrum radio communication devices shown in Figs. 1-14 may be combined in multiple-access embodiments of the present invention that separate simultaneous independent channels by space, frequency and/or code, e.g., space division multiple access (SDMA), frequency division multiple access (FDMA), and code division multiple access (CDMA).

In SDMA embodiments, antenna arrays are used that can be selectively directed in space, e.g., to establish a minimum of two zones. Each transmitter and receiver pair in a zone tunes its corresponding antenna array to embrace only the other one in its transmitter-receiver pair and to exclude other pairs in other zones that represent the other multiple access channels. Embodiments of the present invention distinguish themselves by combining SDMA techniques with stacked-carrier spread spectrum techniques. For more on SDMA design, see, Forssen, et al., "Adaptive Antenna Arrays for GSM900/DSC1800," March 1994, IEEE #0-7803-1927, pp. 605-609. And see, Talwar, et al., "Reception of Multiple Co-Channel Digital Signals using Antenna Arrays with Applications to PCS," 1994, IEEE #0-7803-1825, pp. 790-794. See further, Weis, et al., "A Novel Algorithm For Flexible Beam Forming for Adaptive Space Division Multiple Access Systems," IEEE PIMRC '94, pp. 729a-729e. The combination of CDMA with antenna arrays is addressed by, Naguib, et al., in "Performance of CDMA Cellular

Networks With Base-Station Antenna Arrays: The Downlink," 1994 IEEE, #0-7803-1825, pp. 795-799. And see, Xu, et al., "Experimental Studies of Space-Division-Multiple-Access Schemes for Spectral Efficient Wireless Communications," 1994 IEEE, #0-7803-1825, pp. 800-804. See further, M. Tangemann, "Influence of the User Mobility on the Spatial Multiplex Gain of an Adaptive SDMA System," IEEE PIMRC '94, pp. 745-749.

In FDMA embodiments, subsets of the multiple carriers are used for each channel, e.g., a minimum of two subsets each having a minimum of two frequency diverse carriers to establish a minimum of two channels. Each transmitter and receiver pair in a zone tunes its corresponding carrier subset to exclude other carrier subsets that represent the other multiple access channels. Embodiments of the present invention distinguish themselves by combining FDMA techniques with stacked-carrier spread spectrum techniques.

In CDMA embodiments, several spreading and despreading weights are used, one set for each channel. Such multiple access is used by navigation receivers in the global positioning system (GPS). Embodiments of the present invention distinguish themselves over the prior art by combining CDMA techniques with the stacked-carrier spread spectrum techniques illustrated in Figs. 1-14. For more on CDMA design in a multi-carrier environment, see, Fettweis, et al., "On Multi-Carrier Code Division Multiple Access (MC-CDMA) Modem Design," 1994 IEEE #0-7803-1927, pp. 1670-1674. And see, DaSilva, et al., "Multicarrier Orthogonal CDMA Signals for Quasi-Synchronous Communication Systems," IEEE Journal on Selected Areas in Communication, Vol. 12, No. 5, June 1994. And also, Reiners, et al., "Multicarrier Transmission Technique in Cellular Mobile Communications Systems," March 1994, IEEE #0-7803-1927, pp. 1645-1649. Further see, Yee, et al., "Multi-Carrier CDMA in Indoor Wireless Radio Networks," IEEE Trans. Comm., Vol. E77-B, No. 7, July 1994, pp. 900-904. Using CDMA in the presence of fading channels is addressed by, Stefan Kaiser, "On the Performance of Different Detection Techniques for OFDM-CDMA in Fading Channels," Institute for Communication Technology, German Aerospace Research Establishment (DLR), Oberpfaffenhofen, Germany, 1994. And see, Chandler, et al., "An ATM-CDMA Air Interface For Mobile Personal Communications," IEEE PIMRC '94, pp. 110-113. And further writing on this technology is, Chouly, et al., in "Orthogonal multicarrier techniques applied to direct sequence spread spectrum CDMA systems," 1993 IEEE, #0-7803-0917, pp. 1723-1728.

A combination of multi-carrier CDMA and decorrelating interference cancellation is described by, Bar-Ness, et al., in "Synchronous Multi-User Multi-Carrier CDMA Communication System With Decorrelating Interference Canceller," IEEE PIMRC '94, pp. 184-188.

A multiple access method for stacked-carrier spread spectrum radio communication comprises constructing at a transmitter stacked-carrier spreading gain from the complex amplitude and phase gain of a complex sinusoid for each of a plurality of discrete frequency channels. Then, spreading at the transmitter an arbitrary narrow-band baseband data with a vector multiplier and an inverse frequency channelizer. The next step is to simultaneously transmit from the transmitter the data after spreading over the plurality of discrete frequency channels with the stacked-carrier spreading gain. The receiver despreads the plurality of discrete frequency channels with a vector inner product linear combiner and frequency channelizer, so the arbitrary narrow-band baseband pre-spread data is recovered with relative immunity to channel interference. The frequency channels may be non-contiguous and distributed within multiple bands. Alternatively, the transmitting is such that the frequency channels overlap and include orthogonal frequency-division multiplex-like modulation formats. And alternatively, the transmitting of the data is in packets, wherein baseband data is spread, transmitted, and despread in discrete packets in an orthogonal frequency-division multiplex-like-based frequency channelizer structure.

Packets may be overlapped, contiguous, or noncontiguous in time. The preferred embodiment sequentially transmits one or more packets after sequentially receiving one or more packets from the other end of the link. Sequential transmission and reception of multiple packets can allow asymmetric communications, e.g. by transporting more packets in one direction than the other, and can provide increased guard time between transmit and receive mode, for example to combat base-to-base interference problems in cellular communication networks.

Combining discrete multiple tone orthogonal frequency-division multiplex-like and antenna array processing techniques with discrete multiple tone-stacked carrier and antenna array processing techniques takes advantage of dispersion-free properties of discrete multiple tones and discrete multiple tone-stacked carriers. A significant improvement in the performance of adaptive antenna arrays in any application requiring spatial interference cancellation is possible by eliminating the need to mitigate stationary and quasi-stationary linear dispersion ahead of the adaptive receiver (e.g., due to front-end receiver imperfections, non-zero array apertures, and fixed multipath scatterers and reflectors). This is particularly useful in cellular point-to-multipoint communication networks including space division multiple access (SDMA) topologies for communicating between multiple users on the same set of frequency channels, since each spatial processor must form (potentially) deep nulls in the direction of the interfering users within that cell.

Code division multiple-access (CDMA) transmits multiple signals over the same set of frequency channels using linearly independent (typically orthogonal) sets of spreading gains. These codes are separated at a despreader using appropriate combiner weights.

Direct-sequence spread spectrum systems can benefit from space-division multiple-access type multiple access, interference excision, and channel equalization capability (code nulling technique). Code nulling has been applied to modulation-on-symbol direct-sequence spread spectrum (MOS-DSSS) or modulation-on-pulse direct-sequence spread spectrum (MOP-DSSS) formats where the period of the spreading gain is exactly equal to an integer number of message symbols (nominally one symbol interval). Code nulling may be usefully combined with stacked carrier modulation formats, e.g., for cancellation of spectrally redundant interferers in an HF/VHF frequency hopping intercept system. In the prior art, general frequency hopping intercept techniques including code nulling interference cancellation have been used with a stacked carrier signal simulating a troposcatter communication link. But this technique is extended by the present invention to point-to-point and point-to-multipoint communications where the intended communicators, as well as the interferers, include stacked-carrier spread spectrum modulation formats. For example, data-directed blind adaptation methods are further included to optimize the despreaders based on known properties of the traffic and pilot data transported by the communication systems.

The present invention combines stacked-carrier spread spectrum-based communications and code nulling based interference cancellation for a communication system with higher capacity, higher tolerance to channel distortion, and less reliance on correlation between spreading gains. Near-orthogonality is not required, and embodiments of the invention have less sensitivity to narrow-band interference or other system member stacked-carrier spread spectrum signals. Such effects are optimized when code nulling based interference cancellation is combined in a stacked-carrier spread spectrum communication network. In particular, a stacked-carrier spread spectrum communication link including code nulling based interference cancellation can support twice as many links as the equivalent modulation-on-symbol-direct-sequence spread spectrum system, given the same spreading gain and code nuller (linear combiner) complexity.

The invention combines code nulling based interference cancellation and data-directed methods used to adapt the despreaders in the network. Such combination provides a system with significant advantages over competing methods for point-to-point and point-to-multipoint (multiple access) communications. Such systems can take advantage of a full time-bandwidth product of the communication system, thereby reducing the acquisition and tracking time of the despreaders in the system. Such systems also despread and demodulate intended stacked-carrier

spread spectrum signals of interest to the despreader, without knowledge of the spreading gains included at the signal transmitters (blind despreading property), thereby simplifying or eliminating the code selection strategy used in the network, and allowing the use of retrodirective techniques that optimize the spreading gain based on the communication channel and network. Null interfering (in-cell or out-of-cell) stacked-carrier spread spectrum signals are received by the despreader, without knowledge of the spreading gains or content of the interfering signals, thereby providing significant complexity improvements over (typically nonlinear) sequential methods that must demodulate and re-modulate the interferers as well as the signals of interest to the receiver. Automatic compensation is provided for stationary linear channel dispersion, including dispersive effects induced within the system front-end, without knowledge or actual estimation of the channel dispersion, thereby reducing the complexity of the despreading methods as well as the system hardware.

Code nulling extends to spatial processing techniques, and facilitates the use of retrodirective transmission methods to greatly improve the performance and cost efficiency of the overall network.

Combining code nulling and spatial processing techniques with adaptive antenna arrays for beam-steering improves the range of an otherwise conventional communication transceiver. The combination can also increase the capacity of a multicell network by reducing the interference presented to adjacent cells. The null-steering for interference cancellation improves the capacity of a communication network by allowing tighter packing. The tighter packing is made possible by separating the frequency-coincident in-cell users in space-division multiple access topologies. Antenna arrays can be combined with code nulling techniques in a straightforward manner by increasing the code-nuller dimensions to combine spatial channels as well as time channels, e.g., in MOS-DSSS systems, or by increasing the code-nuller dimensions to combine spatial channels as well as frequency channels, e.g., in stacked-carrier spread spectrum systems.

The stacked-carrier spread spectrum modulation format allows a despreader to reduce the stacked-carrier spread spectrum spreading gain as the number of spatial channels grows, in order to keep the complexity of the code nuller constant as a function of the number of antenna elements. This provides constant data-directed receiver adaptation time. The linear complexity growths as the number of antenna and users in the communication link grows. And, the spatial distribution of users is reduced as the number of antenna beams grows.

The combination of code nulling data-directed-adaption retrodirective transmission techniques and stacked-carrier spread spectrum modulation provides a superior communications mode. Point-to-point, as well as point-to-multipoint communication links, have increased user

capacity, range, power and/or cost efficiencies that exceed those for full channel preemphasis methods.

Combining stacked-carrier spread spectrum and adaptive antenna array processing helps to remove spatially coherent interference, e.g., in cellular stacked-carrier spread spectrum networks where the interferers can be other member signals in the network, and where multi-element antenna arrays are used mainly at the base stations in the network.

In Fig. 17 a time-frequency format example of a time-division duplex communication system is shown.

Fig. 18 shows an active tone format of a basic DMT modem.

In Fig. 19 a transmit/receive calibration method is shown. There are two separate modes for system calibration and compensation. SCSS cal signal injected into a receiver at cal switch, measures the receive path dispersion. SCSS cal signal routed through a transmit modulator to the output receiver through transfer switch measures combined transmit and receive path dispersion. Transmit path is derived from combined receive and transmit cal data. Compensation is performed in DSP back-end by transmitting and processing SCSS cal waveform.

Fig. 20 diagrams an integrated single-antenna T/R and DMT modem (DMT based SCMA).

Fig. 21 is a diagram of a general example single-link code-gated cross-SCORE spreading operation. It is a preferred mode for single-link processing. It allows the use of a cross-SCORE algorithm with the fastest convergence time (lowest TBP). It is unaffected by timing and Doppler offset. It can reliably remove K_{array} interferers within each cell. It can separate K_{array} SCSS signals. Its deficiencies are that it cannot reliably separate > K_{array} SCSS signals (code nulling not performed) and it misadjusts relative to max-SINR solution in highly frequency-variant environments.

Fig. 22 is an example of a single-link code-gated cross-SCORE despreading operation with K_{spread} cell subsets.

Fig. 23 is an example of a single-link cross-SCORE algorithm with N_{frame} packets/adapt frame. Despreader weights are computed from the dominant mode of multiset cross-SCORE eigenequation.

Fig. 24 is an example of single adapt frame autocorrelation statistics computation.

Fig. 25 is an example of cross-SCORE eigenequation with K_{spread} cell subsets. Despreader weights are computed from the dominant mode of multiset cross-SCORE eigenequation.

Fig. 26 is an example of a code key generator with $K_{part} < K_{spread}$ cell subsets.

Fig. 27 is an example of an equivalent code key applicator with $K_{part} < K_{spread}$ cell subsets.

Fig. 28 is an example of cross-SCORE eigenequation with K_{part} cell subsets. Despreader weights are computed from dominant mode of multiset cross-SCORE eigenequation.

Fig. 29 is an example of cross-SCORE eigenequation with two cell subsets. Despreader weights are computed from dominant mode of multiset cross-SCORE eigenequation.

Fig. 30 is an example of a multi-link code-gated cross-SCORE spreader. It is an improved mode for multi-link processing. It allows tailoring of cross-SCORE convergence time to SCSS interference conditions. It is unaffected by timing and Doppler offset. It can reliably remove K_{array} interferers within each cell. It can separate $K_{array} \cdot K_{score}$ SCSS signals. Its deficiencies are that it cannot reliably separate $> K_{array} \cdot K_{score}$ SCSS signals (incomplete code nulling) and it misadjusts relative to max-STAR solution in highly frequency-variant environments.

Fig. 31 is an example of single-link code-gated auto-SCORE spreading operation with gating over frequency and two cell subsets. It is a preferred mode for high-mobility systems. It can separate $K_{array} \cdot K_{score}$ SCSS links. It can remove K_{array} non-SCSS interferers within each cell. It is unaffected by timing and Doppler offset. Its deficiencies are it cannot separate $> K_{score}$ SCSS links and requires (simple) timing tracking as part of the despreading algorithm.

Fig. 32 is an example of single-link code-gated auto-SCORE despreading operation with gating over frequency and two cell subsets.

Fig. 33 is an example of auto-SCORE eigenequation with gating over frequency and two cell subsets.

Fig. 34 is an example of single-link code-gated auto-SCORE spreading with gating over time and a rate one-half of redundancy gate. It is the preferred mode for low-mobility systems. It can separate $K_{array} \cdot K_{spread}$ SCSS links. It can remove K_{array} non-SCSS interferers within each cell. It is unaffected by timing and Doppler offset. Three dB SNR gain is provided at the despreader. Its deficiencies are that it cuts capacity in half and requires (simple) Doppler tracking as part of the despreading algorithm.

Fig. 35 is an example of single-link code-gated auto-SCORE despreading with gating over time and a rate one-half of redundancy gate.

In summary, adaptive antenna arrays can be used to increase network system capacity by beam-steering, null-steering, or combined beam-and-null steering. Such null steering or combined beam and null steering technologies are combined in this invention with DMT/OFDM frequency channelizers outside of the channelizers use as SCSS spreaders/despreaders.

Although the present invention has been described in terms of the presently preferred embodiments, it is to be understood that the disclosure is not to be interpreted as limiting. Various alterations and modifications will no doubt become apparent to those skilled in the art

after having read the above disclosure. Accordingly, it is intended that the appended claims are interpreted as covering all alterations and modifications as fall within the true spirit and scope of the invention.

What is claimed is:

1. A multiple access communication system, comprising:

at least one radio transmitter for transmitting a plurality of radio frequency (RF) carriers;

at least one radio receiver for receiving said at least one subset of at least two of said plurality of radio frequency carriers;

at least one spreader connected to each of the transmitters for independently and redundantly modulating the amplitude and phase of at least two of said (RF) carriers with a first digital spreading gain and first data;

at least one despreader connected to each of the receivers for independently demodulating the amplitude and phase of at least two of said (RF) carriers with said first digital spreading gain to recover said first data; and

multiple access means connected to the transmitters, receivers, spreaders and despreaders for providing separate channels of communication by at least one of space-division multiple access (SDMA), frequency-division multiple access (FDMA) and code-division multiple access (CDMA).

2. The system of claim 1, wherein:

said SDMA further comprises an antenna array connected to said transmitters and receivers, and provides for selective data channel transmission and reception between pairs of transmitters and receivers according to their relative spatial positions.

3. The system of claim 1, wherein:

said FDMA further comprises a minimum number of radio frequency carriers to provide for at least two subsets of carriers for communicating additional data channels between pairs of transmitters and receivers according to matching subsets of said radio frequency carriers.

4. The system of claim 1, wherein:

said CDMA comprises including at least a second digital spreading gain and second data, and provides for communication of at least said first and second data between pairs of transmitters and receivers according to matching subsets of said digital spreading gains.

5. A radio transmitter system providing for spatial and frequency dispersion, comprising:

an antenna system having "n" positive-integer number of individual antennas spatially distributed in a group and providing for as many as "n" positive-integer number of spatially distributed communication channels;

a radio frequency amplifier bank having individual amplifiers connected to corresponding antennas of the antenna system wherein each amplifier has an adjustable gain to provide for a controlled steering of beams and nulls of radio frequency signal transmissions to said radially-distributed spatial communication channels;

a discrete multitone stacked carrier spread spectrum transmit modulator connected to the amplifier bank and providing for a frequency-dispersed plurality of communication channels;

a spreader having respective outputs connected to the modulator and a data input and providing for a spreading of data concurrently across all of said frequency-dispersed plurality of communication channels; and

steering means connected to said radio frequency amplifier bank and providing for a selection of one of said "n" positive-integer number of radially-distributed spatial communication channels.

6. The system of Claim 5, said system further providing for polarization dispersions wherein said individual antennas are further distributed in polarization.

7. A radio receiver system providing for spatial and frequency dispersion, comprising:

an antenna system having "n" positive-integer number of individual antennas spatially distributed in a group and providing for as many as "n" positive-integer number of radially-distributed spatial communication channels;

a radio frequency amplifier bank having individual amplifiers connected to corresponding antennas of the antenna system and each amplifier having an adjustable gain which provides for a controlled steering of beams and nulls of radio frequency signal transmissions to said radially-distributed spatial communication channels;

a discrete multitone stacked carrier spread spectrum demodulator connected to the amplifier bank and providing for a frequency-dispersed plurality of communication channels;

a despreader having respective inputs connected to the demodulator and a data output and providing for a despreading of data included across all of said frequency-dispersed plurality of communication charnels; and

steering means connected to said radio frequency amplifier bank and providing for a selection of one of said "n" positive-integer number of radially-distributed spatial communication channels.

8. The system of claim 6, further comprising:

computational means for determining a radially-distributed spatial communication channel address of a target receiver.

9. A multiple-access method for stacked-carrier spread spectrum radio communication, comprising the steps of:

constructing at a transmitter a stacked-carrier spreading gain from the complex amplitude and phase gain of a complex sinusoid for each of a plurality of discrete frequency communication channels;

spreading at said transmitter an arbitrary narrow-band baseband pre-spread data with a vector multiplier and an inverse frequency communication channelizer;

simultaneously transmitting from said transmitter said data after spreading over said plurality of discrete frequency communication channels with said stacked-carrier spreading gain; and

despreading said plurality of discrete frequency communication channels at a receiver with a vector inner product linear combiner and frequency communication channelizer, wherein said arbitrary narrow-band baseband pre-spread data is recovered with relative immunity to communication channel interference.

10. The method of claim 9, wherein:

said constructing is such that said frequency communication channels are non-contiguous and distributed within multiple bands.

11. The method of claim 10, wherein:

said simultaneously transmitting is such that said frequency communication channels overlap and include orthogonal frequency-division multiplex-like modulation formats.

12. The method of claim 11, wherein:

said simultaneously transmitting is such that said data is packetized, wherein baseband data is spread, transmitted, and despread in discrete packets in an orthogonal frequency-division

multiplex-like-based frequency communication channelizer structure, said packets being overlapped, contiguous, or non-contiguous in time.

13. An interference cancellation method for stacked-carrier spread spectrum radio communication, comprising the steps of:

constructing at a transmitter a stacked-carrier spreading gain from the complex amplitude and phase gain of a complex sinusoid for each of a plurality of discrete frequency communication channels;

spreading at said transmitter an arbitrary narrow-band baseband pre-spread data with a vector multiplier and an inverse frequency communication channelizer;

simultaneously transmitting from said transmitter said data after spreading over said plurality of discrete frequency communication channels with said stacked-carrier spreading gain;

despreading said plurality of discrete frequency communication channels at a receiver with a vector inner product linear combiner and frequency communication channelizer, wherein said arbitrary narrow-band baseband pre-spread data is recovered with relative immunity to communication channel interference; and

code nulling with interference cancellation means according to information obtained from the step of despreading.

14. An adaptive antenna array method for stacked-carrier spread spectrum radio communication, comprising the steps of:

constructing at a transmitter a stacked-carrier spreading gain from the complex amplitude and phase gain of a complex sinusoid for each of a plurality of discrete frequency communication channels;

spreading at said transmitter an arbitrary narrow-band baseband pre-spread data with a vector multiplier and an inverse frequency communication channelizer;

simultaneously transmitting from said transmitter said data after spreading over said plurality of discrete frequency communication channels with said stacked-carrier spreading gain;

despreading said plurality of discrete frequency communication channels at a receiver with a vector inner product linear combiner and frequency communication channelizer, wherein said arbitrary narrow-band baseband pre-spread data is recovered with relative immunity to communication channel interference; and

adjusting the gains of several amplifiers connected to an antenna array and said transmitter according to said despread baseband data.

15. A time-division duplex method for stacked-carrier spread spectrum radio communication, comprising the steps of:

dividing time into time slots reserved for transmitting a first plurality of discrete frequency communication channels to a distant receiver and for receiving a second plurality of discrete frequency communication channels from a distant transmitter;

constructing at a proximal transmitter a stacked-carrier spreading gain from the complex amplitude and phase gain of a complex sinusoid for at least one of said first or second plurality of discrete frequency communication channels;

spreading at said proximal transmitter an arbitrary narrow-band baseband pre-spread data with a vector multiplier and an inverse frequency communication channelizer;

simultaneously transmitting from said proximal transmitter said data after spreading over said first plurality of discrete frequency communication channels with said stacked-carrier spreading gain;

despreading said second plurality of discrete frequency communication channels at a proximal receiver with a vector inner product linear combiner and frequency communication channelizer, wherein said arbitrary narrow-band baseband pre-spread data is recovered with relative immunity to communication channel interference; and

controlling the step of dividing time with precision time information obtained from a timing signal source that is available to both said distant and proximal transmitters and both said distant and proximal receivers.

16. The method of claim 15, wherein:

the step of controlling includes receiving system time information from orbiting navigation satellites.

17. A radio transmitter system, comprising:

a multi-tone transmitter array for multi-spectral carrier signal transmission according to a spectral weight calculation by a computer that discriminates between a first plurality of individual remote receivers; and

an antenna array connected to the transmitter array and that provides for spatial adjustments of the transmitted power of said multi-spectral carrier signal transmission according to a spatial weight calculation by a computer that discriminates between a second plurality individual remote receivers.

18. A radio receiver system, comprising:

a multi-tone receiver array for multi-spectral carrier signal reception according to a spectral weight calculation by a computer that discriminates between a first plurality of individual remote transmitters; and

an antenna array connected to the receiver array and that provides for spatial adjustments of the received power of said multi-spectral carrier signal reception according to a spatial weight calculation by a computer that discriminates between a second plurality individual remote transmitters.

19. A radio communication system, comprising:

a multi-tone transmitter array for multi-spectral carrier signal transmission according to a spectral weight calculation by a computer that discriminates between a first plurality of individual remote receivers;

a first antenna array connected to the transmitter array and that provides for spatial adjustments of the transmitted power of said multi-spectral carrier signal transmission according to a spatial weight calculation by a computer that discriminates between a second plurality individual remote receivers;

a multi-tone receiver array for multi-spectral carrier signal reception according to a spectral weight calculation by a computer that discriminates between a first plurality of individual remote transmitters; and

a second antenna array connected to the receiver array and that provides for spatial adjustments of the received power of said multi-spectral carrier signal reception according to a spatial weight calculation by a computer that discriminates between a second plurality individual remote transmitters.

20. The system of claim 19, further comprising:

a computer connected to the transmitter array and first antenna array that provides for a single omnibus calculation of both the spectral weight and spatial weights.

21. The system of claim 19, further comprising:

a computer connected to the receiver array and second antenna array that provides for a single omnibus calculation of both the spectral weight and spatial weights.

22. The system of claim 19, further comprising:

a computer connected to the transmitter array, the receiver array, the first antenna array, and the second antenna array and that provides for a single omnibus calculation of both the spectral weight and spatial weights for use by the receiver array and second antenna array that repeats both the spectral weight and spatial weights for use by the transmitter array and first antenna array;

wherein a particular spatial and spectral characteristic of said first and second plurality individual remote transmitters is exploited to optimize a return transmission to said first and second plurality individual remote receivers.

23. A method of recovering a digital communication signal that was spread and modulated onto each of a plurality of stacked carrier signals using a distinct spreading gain for each of said plurality of stacked carrier signals, transmitted across a wireless medium, and received at a receiver as a plurality of received stacked carrier signals, comprising the steps of:

channelizing each of said received stacked carrier signals to identify a baseband signal for each of said received stacked-carrier signals, said received stacked carrier signals having a channel bandwidth that is separable from said channel bandwidth of other of said plurality of received stacked carrier signals;

despreading by applying despread weights that are different than said spreading gains to each of said received baseband signals and combining said received baseband signals to obtain a baseband signal that compensates for interference and maximizes a signal to noise and interference ratio; and

removing at least one of time distortion and frequency distortion that exists in said baseband signal to obtain a recovered digital communication signal that corresponds to said digital communication signal.

24. A method according to claim 23 wherein said step of despreading blindly despreads said plurality of received stacked carrier signals.

25. A method according to claim 24 wherein said blind despreading uses a dominant mode of a generalized eigenequation.

26. A method according to claim 25 wherein said blind despreading uses dominant modes of a generalized eigenequation.

27. A method according to claim 26 wherein said eigenequation is a code gated self coherence restoral eigenequation.

28. A method according to claim 25 wherein a maximum code gated self restoral eigenvalue of said generalized eigenequation is decremented by a predicted mean and scaled by a predicted standard deviation during said step of blind despreading.

29. A method according to claim 23 wherein said steps of channelizing, despreading, and removing are each repeatedly performed on each of a sequential plurality of received stacked carrier signals to obtain a recovered sequential plurality of related digital communication signals.

30. A method according to claim 29 wherein said sequential plurality of received stacked carrier signals are asynchronous.

31. A method according to claim 29 wherein each of said sequential plurality of received stacked carrier signals are received during associated time division duplex intervals and a network clock is used to determine said time division duplex intervals.

32. A method according to claim 29 wherein receive time duplex intervals are asymmetric with respect to transmit time duplex intervals.

33. A method according to claim 31 wherein receive time duplex intervals are symmetric with respect to transmit time duplex intervals.

34. A method according to claim 31 wherein said sequential plurality of received stacked carrier signals are received as a plurality of packets during a single time division duplex interval.

35. A method according to claim 31 wherein said sequential plurality of received stacked carrier signals are received as a single packet during a single time division duplex interval.

36. A method according to claim 31 wherein said network clock is derived from a universal time.

37. A method according to claim 31 wherein said network clock is derived from data within said digital communication signal.

38. A method according to claim 23 wherein second spreading gains for a second data communication signal transmitted as a second plurality of stacked carrier signals from said receiver are adaptively determined based upon said despread weights so that minimum radiation is directed at interfering frequencies.

39. A method according to claim 38 wherein said second spreading gains are set proportional to conjugated despread weights so that a gain pattern of said second plurality of stacked carrier signals is substantially the same as the gain pattern of said plurality of stacked carrier signals.

40. A method according to claim 23 wherein a plurality of recovered digital communication signals are simultaneously recovered from a plurality of received stacked carrier signals, each of said plurality of digital communication signals having an associated distinct code key that is used during said step of despreading to discriminate each of said plurality of received stacked carrier signals.

41. A method according to claim 40 wherein each of said distinct code keys modulates only some of said plurality of received stacked carrier signals.

42. A method according to claim 41 wherein each of said distinct code keys modulates one of even and odd received stacked carrier signals from said plurality of received stacked carrier signals.

43. A method according to claim 23 wherein said digital communication signal is a plurality of symbols, and each of said symbols is modulated onto each of said plurality of stacked carrier signals at a different discrete tone.

44. A method according to claim 23 wherein said digital communication signal is a plurality of bits, and each of said bits is modulated onto each of said plurality of stacked carrier signals at a different discrete tone.

45. A method according to claim 23 wherein said time distortion is due to a doppler frequency offset.

46. A method according to claim 23 wherein said frequency distortion is due to time dispersion.

47. A method according to claim 23 wherein said frequency distortion is a propagation delay.

48. A method according to claim 23 wherein said step of removing removes both time distortion and frequency distortion.

49. A method according to claim 48 wherein said time distortion is due to a doppler frequency offset.

50. A method according to claim 48 wherein said frequency distortion is due to time dispersion.

51. A method according to claim 48 wherein said frequency distortion is a propagation delay.

52. A method according to claim 48 wherein said frequency distortion is a propagation delay.

53. A method according to claim 23 wherein said received stacked carrier signals contain a guard time interval to compensate for unknown propagation delay, and said method does not synchronize said received stacked carrier signals until said step of despreading is complete.

54. A method according to claim 23 wherein said received stacked carrier signals contain a guard frequency band to compensate for unknown doppler frequency offset, and said method does not synchronize staid received stacked carrier signals until said step of despreading is complete.

55. A method of recovering a digital communication signal that was spread using spreading gains when transmitted and received at a receiver as received signals comprising the steps of:

despreading said received signals to obtain despread signals; and

removing Doppler time delay from said despread signals to obtain a recovered digital communication signal that corresponds to said transmitted digital communication signal.

56. A method of recovering a plurality of transmitted symbol that was spread using distinct spreading gains, said symbol being received at a receiver as a plurality of discrete multiple tones having substantial frequency diversity, said method of recovering comprising the steps of:

despreading each of said plurality of discrete multiple tones to obtain a plurality of despread multiple tones, each of said plurality of despread multiple tones corresponding to one of said plurality of symbols; and

removing doppler time delay from said despread multiple tones obtain a recovered plurality of symbols corresponding to said transmitted plurality of symbols.

57. A method according to claim 56 wherein said transmitted symbols were spread and modulated onto each of a plurality of stacked carrier signals using a distinct spreading gain for each of said plurality of stacked carrier signals, each of said plurality of stacked carrier signals having a channel bandwidth that is separable from said channel bandwidth of other of said plurality of stacked carrier signals.

58. A method according to claim 57 wherein said step of despreading blindly despreads said plurality of received stacked carrier signals.

59. A method according to claim 58 wherein said blind despreading uses a dominant mode of a generalized eigenequation.

60. A method according to claim 58 wherein said blind despreading uses dominant modes of a generalized eigenequation.

61. A method according to claim 60 wherein said eigenequation is a code gated self coherence restoral eigenequation.

62. A method according to claim 59 wherein a maximum code gated self restoral eigenvalue of said generalized eigenequation is decremented by a predicted mean and scaled by a predicted standard deviation during said step of blind despreading.

63. A method of spreading a digital communication signal comprising the steps of:

spectrally spreading said digital communication signal onto a plurality of stacked carrier signals to obtain a plurality of spectrally spread digital communication signals, each of said

plurality of stacked carrier signals having a channel bandwidth that is separable from said channel bandwidth of other of said plurality of stacked carrier signals;

spatially spreading each of said plurality of spectrally spread digital communication signals to obtain a plurality of spatially and spectrally spread digital communication signals; and

transmitting, from each antennae element of a multi-element antennae array, associated ones of said plurality of spatially and spectrally spread digital communication signals across a wireless medium to a receiver.

64. A method of digitally communicating comprising the steps of:

spreading, in a moving transmitter, digital information that was spread using distinct spreading gains onto each of a plurality of stacked carrier signals, each of said plurality of stacked carrier signals having a channel bandwidth that is separable from said channel bandwidth of other of said plurality of stacked carrier signals;

transmitting each of said stacked carrier signals across a wireless medium from said moving transmitter to a receiver;

receiving, at said receiver, said transmitted plurality of stacked carrier signals as received stacked carrier signals;

channelizing each of said received stacked carrier signals to identify a baseband signal for each of said received stacked carrier signals;

despreading by applying despread weights that are different than said spreading gains to each of said received baseband signal and combining said received baseband signals to obtain a baseband signal that compensates for interference and maximizes a signal to noise and interference ratio; and

processing said baseband signal to obtain recovered digital information that corresponds to said transmitted digital information.

65. A method according to claim 64 wherein said processing step removes Doppler time delay from said baseband signal.

66. A method according to claim 64 wherein:

a plurality of moving transmitters each perform said step of spreading and transmitting said digital information, each moving transmitter spreading said digital information using a random distinct, non-orthogonal spreading gain to spread said digital information onto each of said plurality of stacked carrier signals;

said step of receiving said transmitted plurality of stacked carrier signals receives said transmitted stacked carrier signals from each of said transmitters at a receiver;

said step of channelizing channelizes each of said received stacked carrier signals;

said despreading step combines each of said separately channelized plurality of received stacked carrier signals to obtain, for each of said plurality of transmitters, one baseband signal corresponding to said digital information transmitted by one of said plurality of transmitters; and

said step of processing processes each of said baseband signals to obtain said recovered digital communication signal that corresponds to said transmitted digital communication signal for each of said moving transmitters.

67. A method according to claim 66 wherein said processing step removes Doppler time delay that exists in each of said baseband signals.

68. A method according to claim 67 wherein said plurality of moving transmitters are geographically separated and each present a distinct transmission path to said receiver.

69. A method according to claim 67 wherein a multi-element antenna array is used at said receiver, different ones of said elements are used to differentiate zones, and space division multiple access is employed to differentiate different ones of said transmitters.

70. A method according to claim 65 wherein frequency division multiple access is employed to differentiate different ones of said transmitters.

71. A method according to claim 67 wherein code division multiple access is employed to differentiate different ones of said transmitters.

72. A method according to claim 71 further including the step of modulating a distinct code key onto some of said plurality of stacked carrier signals associated with each of said transmitters.

73. A method of digitally communicating comprising the steps of:

spreading digital information, at a first station having a multi-element antenna array, using distinct spreading gains, onto each of a plurality of stacked carrier signals, each of said plurality of stacked carrier signals having a channel bandwidth that is separable from said channel

bandwidth of other of said plurality of stacked carrier signals and each of said antenna elements being used to transmit different ones of said stacked carrier signals;

transmitting each of said stacked carrier signals across a wireless medium from said first station to a second station;

receiving, at said second station using a second multi-element antenna array, said transmitted plurality of stacked carrier signals as received stacked carrier signals, each of said second antenna elements being used to receive different ones of said received stacked carrier signals;

channelizing each of said received stacked carrier signals to identify a baseband signal for each of said received stacked carrier signals; and

despreading by applying despread weights that are different than said spreading gains to each of said received baseband signal and combining said received baseband signals to obtain a baseband signal that compensates for interference and maximizes a signal to noise and interference ratio.

74. A method according to claim 73 wherein second spreading gains for a second data communication signal transmitted as a second plurality of stacked carrier signals from said multi-element antennae array of said second station are adaptively determined based upon said despread weights so that minimum radiation is directed at interfering frequencies.

75. A method according to claim 74 wherein said second spreading gains are set proportional to conjugated despread weights so that a gain pattern of said second plurality of stacked carrier signals is substantially the same as the gain pattern of said plurality of stacked carrier signals.

76. A method according to claim 73 wherein second spreading gains for a second data communication signal transmitted from said multi-element antennae array of said second station are adaptively determined so that maximum radiation is directed to an intended station.

77. A method according to claim 76 wherein said second spreading gains are adaptively determined using a combined spatial and spectral steering vector.

78. A method according to claim 73 wherein each of said distinct spreading gains associated with each of said plurality of stacked carrier signals is linearly independent and non-orthogonal.

79. A method of digitally communicating comprising the steps of:

spreading, at a first station, digital information onto each of a plurality of stacked carrier signals using a distinct spreading gain for each stacked carrier signal, each of said plurality of stacked carrier signals having a channel bandwidth that is separable from said channel bandwidth of other of said plurality of stacked carrier signals;

transmitting each of said stacked carrier signals across a wireless medium from said first station to a second station;

receiving, at said second station using a multi-element antenna array, said transmitted plurality of stacked carrier signals as received stacked carrier signals;

channelizing on each element of said array each of said received stacked carrier signals to identify a baseband signal for each of said received stacked carrier signals; and

despreading by applying despread weights that are different than said spreading gains to each of said received baseband signal and combining said received baseband signals to obtain a baseband signal that compensates for interference and maximizes a signal to noise and interference ratio, said despreading step using a linear combiner dimensionality which is composed of spectral dimensions and spatial dimensions.

80. A method according to claim 79 wherein second spreading gains for a second data communication signal transmitted as a second plurality of stacked carrier signals from said multi-element antennae array of said second station are adaptively determined based upon said despread weights so that minimum radiation is directed at interfering frequencies.

81. A method according to claim 80 wherein said second spreading gains are set proportional to conjugated despread weights so that a gain pattern of said second plurality of stacked carrier signals is substantially the same as the gain pattern of said plurality of stacked carrier signals.

82. A method according to claim 79 wherein second spreading gains for a second data communication signal transmitted from said multi-element antennae array of said second station are adaptively determined so that maximum radiation is directed to an intended station.

83. A method according to claim 82 wherein said second spreading gains are adaptively determined using a combined spatial and spectral steering vector.

84. A method according to claim 79 wherein said spatial and spectral dimensions is variable for a plurality of different ones of said first stations.

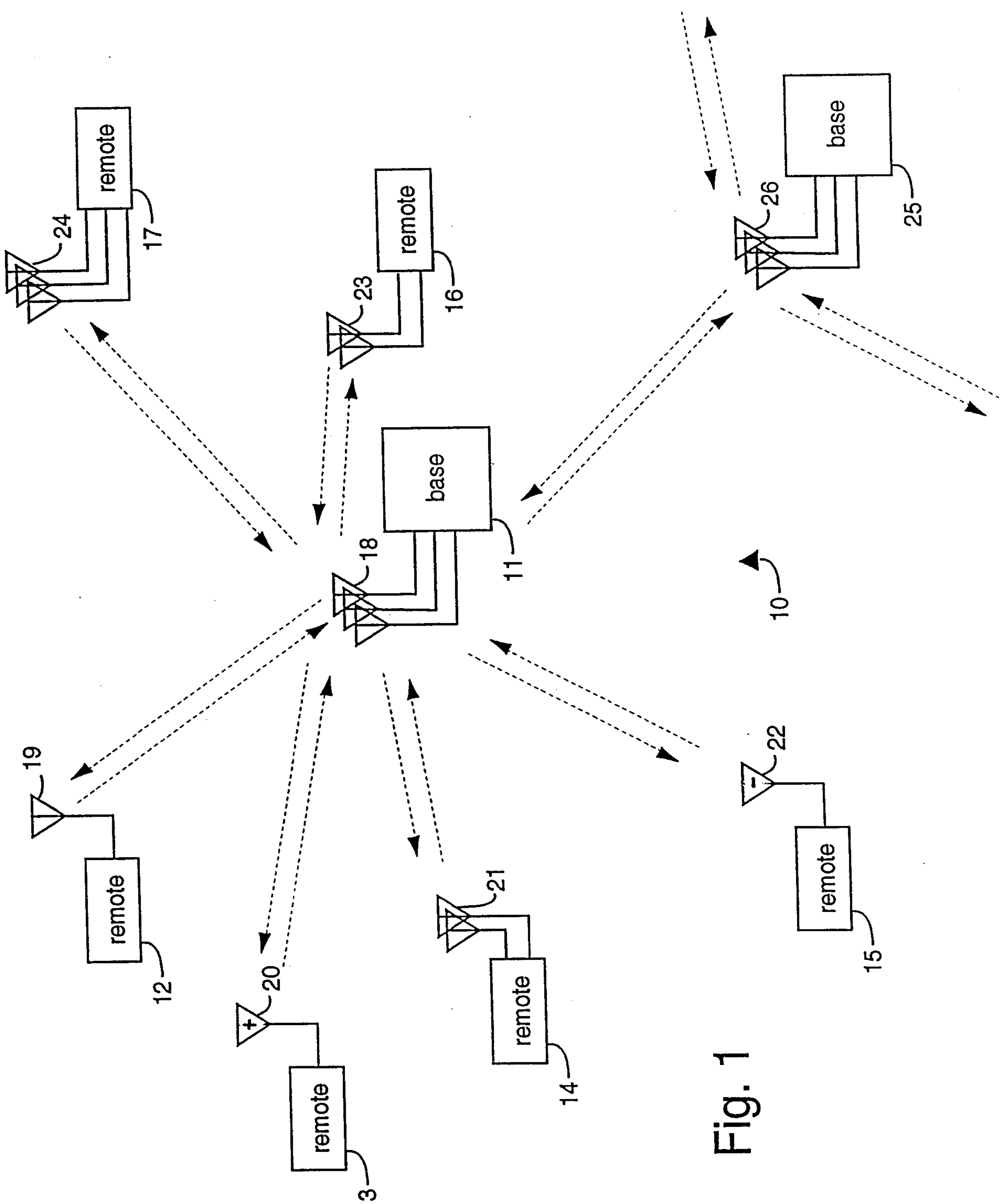
85. A method according to claim 79 wherein each of said distinct spreading gains associated with each of said plurality of stacked carrier signals is linearly independent and non-orthogonal.

86. A method of despreading a digital communication signal received at a receiver from across a wireless medium comprising the steps of:

receiving, at each antennae element of a multi-element antennae array, a plurality of stacked carrier signals, each of said plurality of stacked carrier signals having a channel bandwidth that is separable from said channel bandwidth of other of said plurality of stacked carrier signals;

spatially despreading said plurality of stacked carrier signals to obtain spatially despread digital communication signals; and

spectrally despreading each of said plurality of spatially despread digital communication signals to obtain a spatially and spectrally despread digital communication signal.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base
remote
remote
remote
remote
remote
remote
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11
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Fig. 1

Fig. 2A

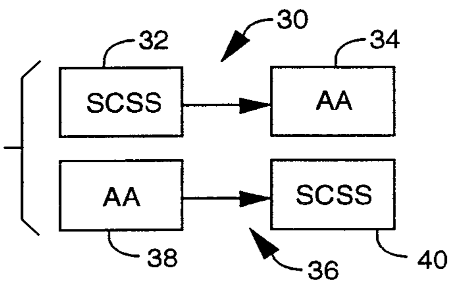


Fig. 2B

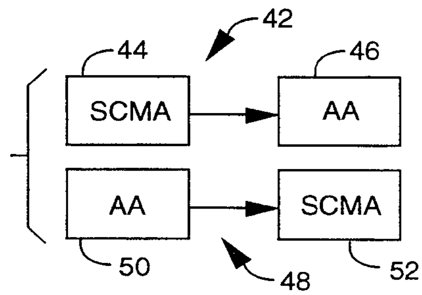


Fig. 3A

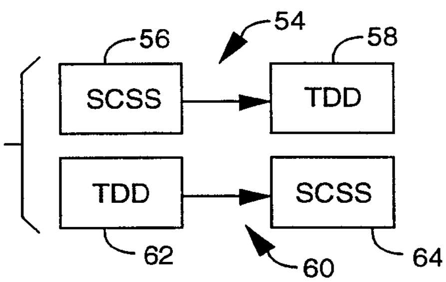


Fig. 3B

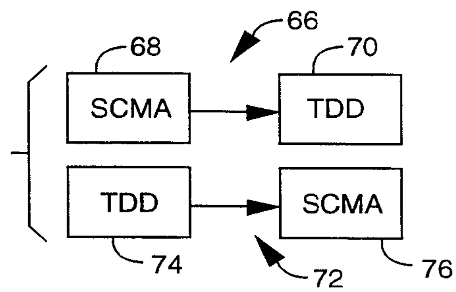


Fig. 4A

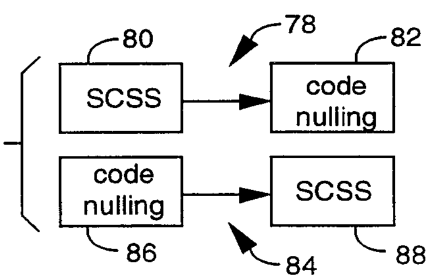


Fig. 4B

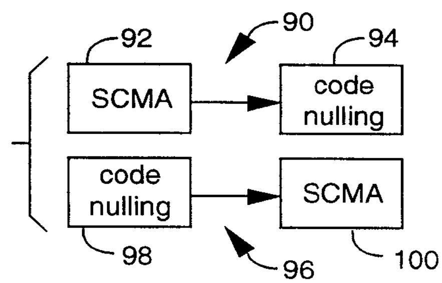


Fig. 5A

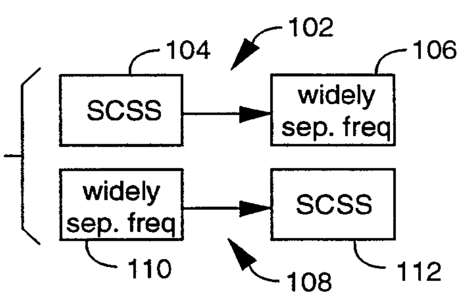


Fig. 5B

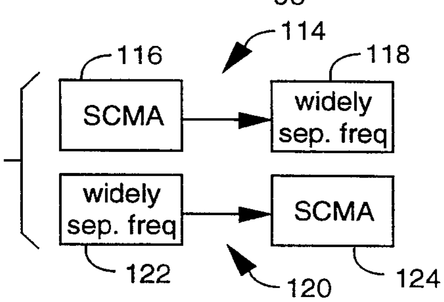


Fig. 6A

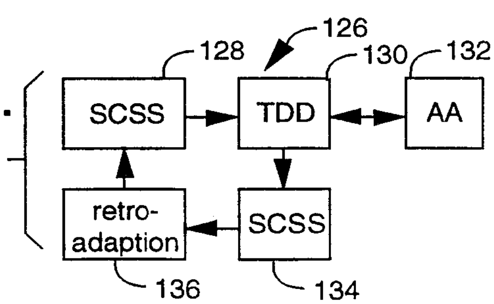
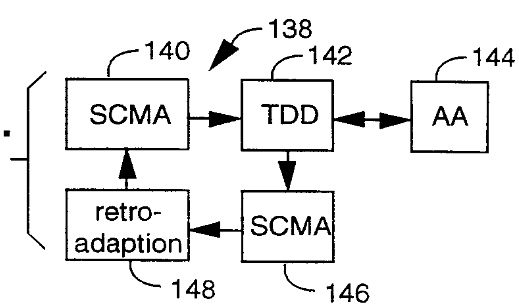
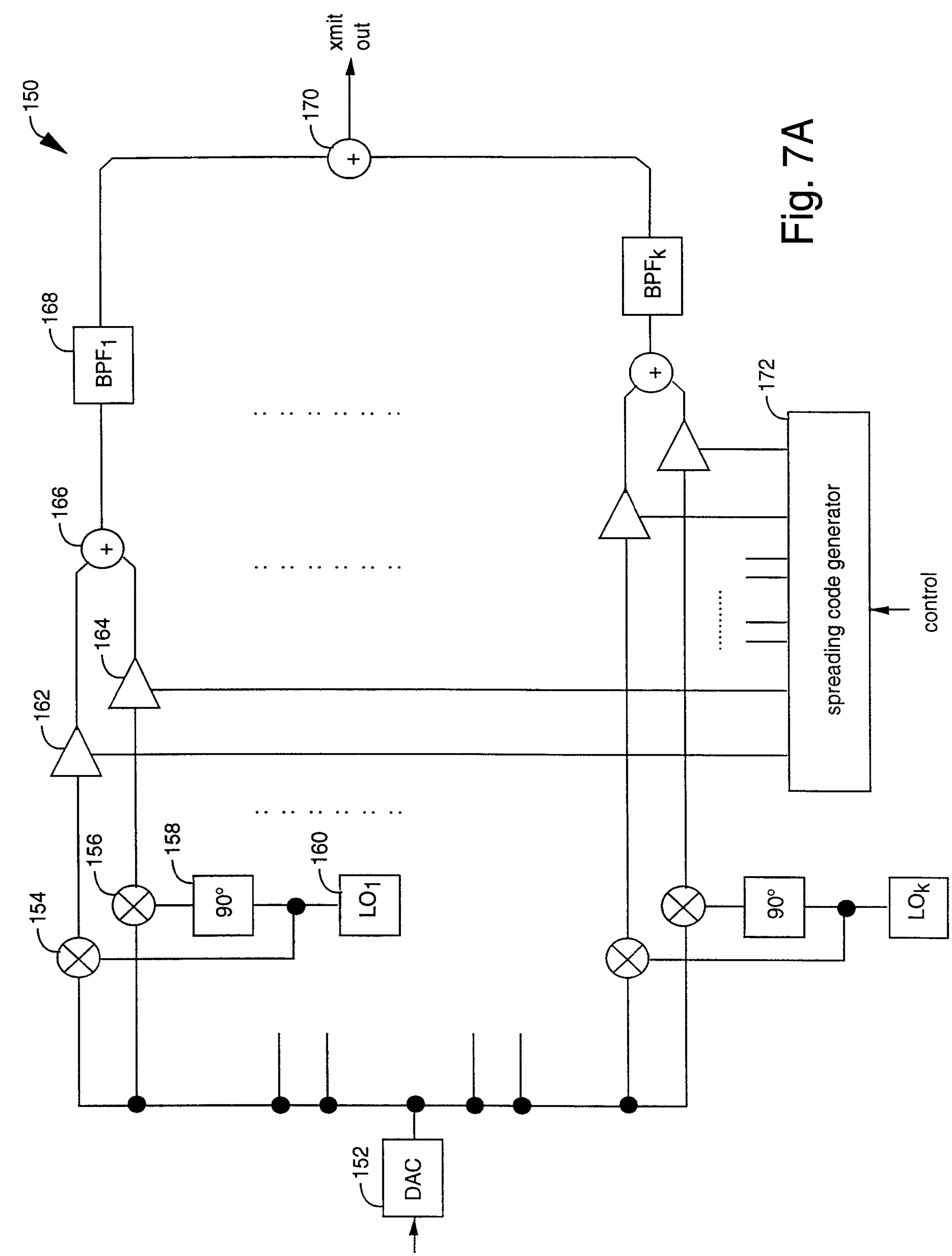


Fig. 6B

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xmit out
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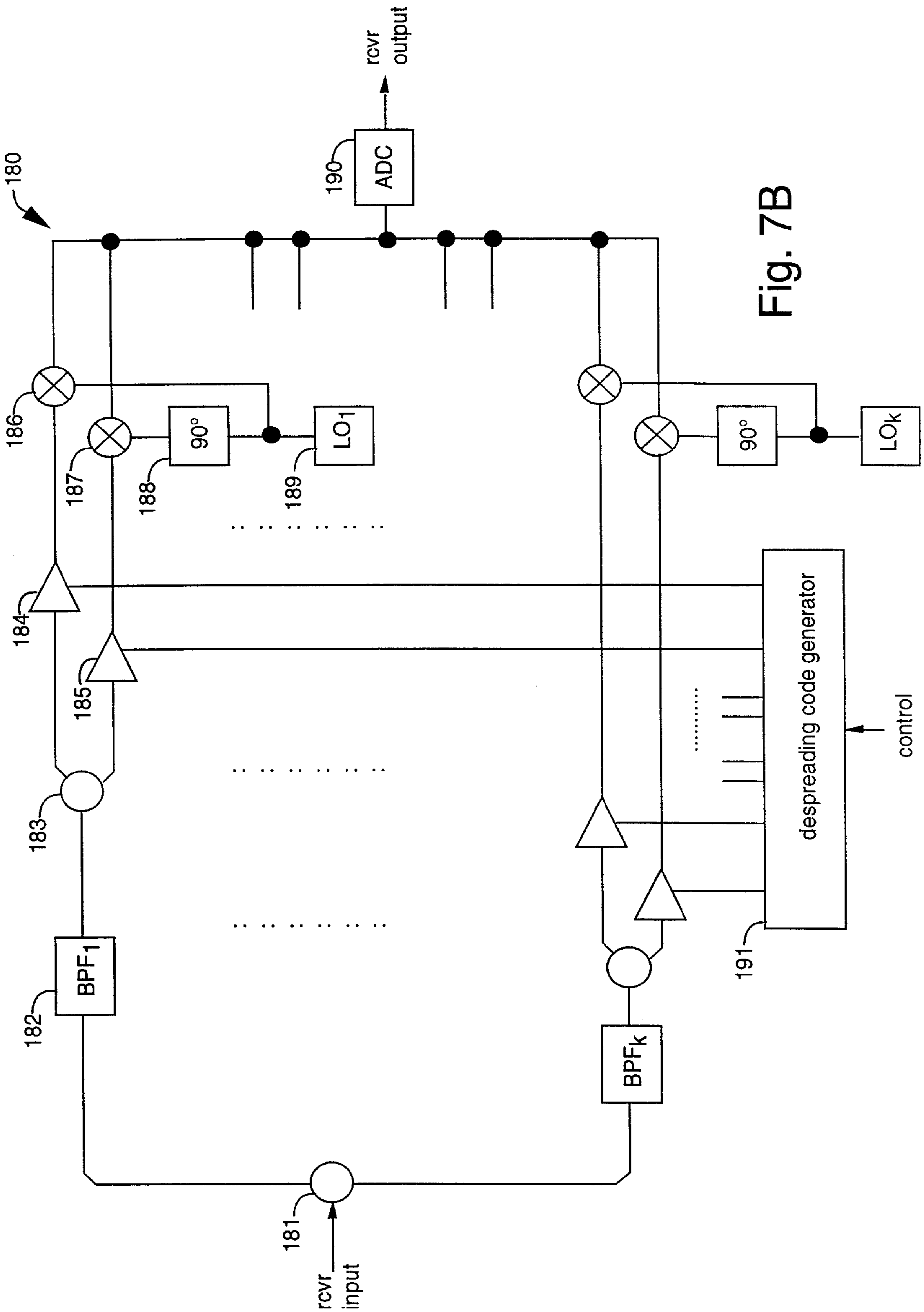
168
BPF1
BPFK
166
164
162
172
spreading code generator
control
154
156
158
160
90°
90°
LO1
LOK
152
DAC

Fig. 7A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ADC
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90°
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LOk
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182
BPF1
BPFk
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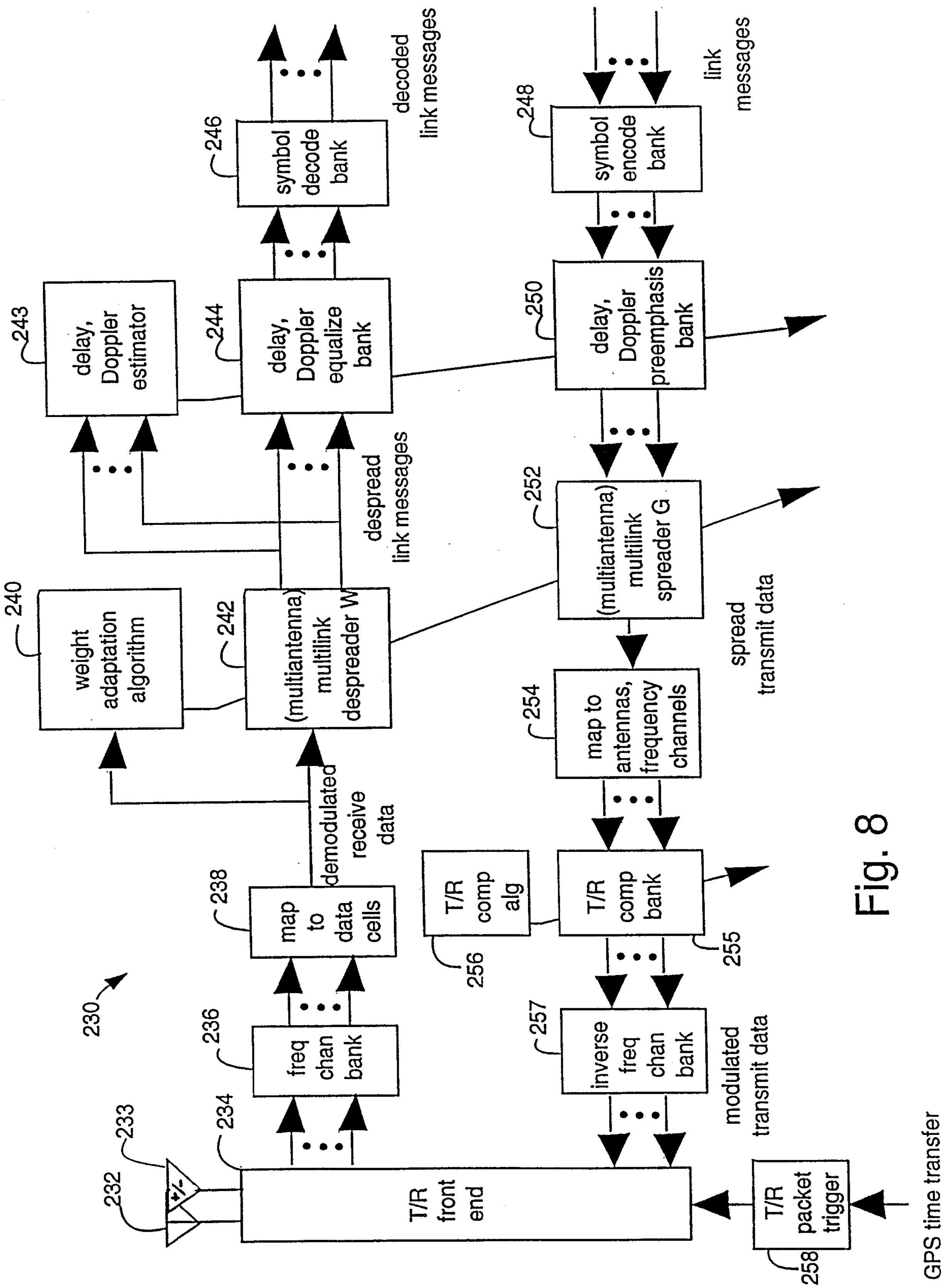
rcvr input
despreading code generator
control
191

Fig. 7B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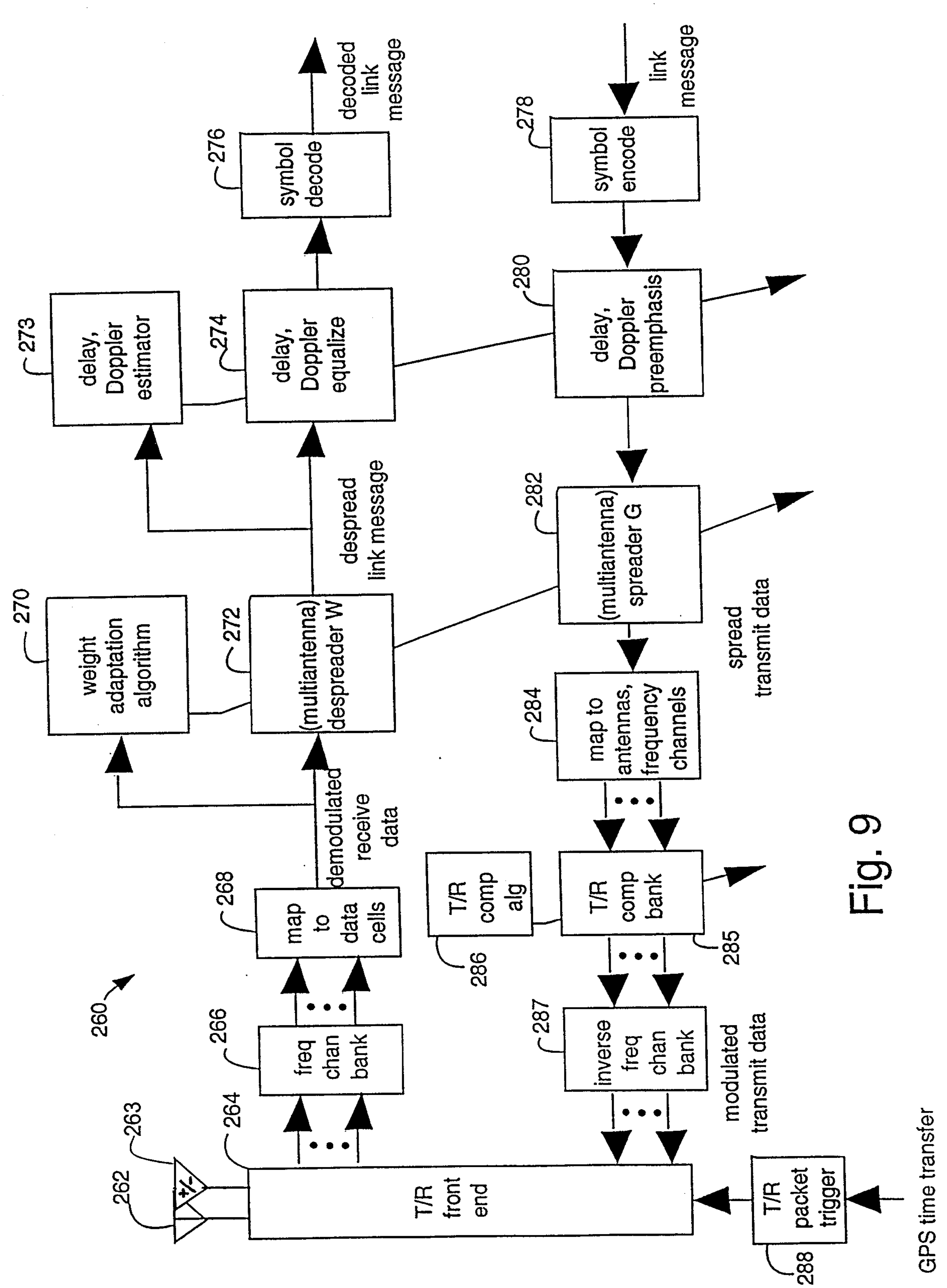
232
233
+/-
234
T/R front end
236
freq chan bank
238
map to data cells
demodulated receive data
240
weight adaptation algorithm
242
(multiantenna) multilink despreader W
despread link messages
243
delay, Doppler estimator
244
delay, Doppler equalize bank
246
symbol decode bank
decoded link messages
248
symbol encode bank
link messages
250
delay, Doppler preemphasis bank
252
(multiantenna) multilink spreader G
spread transmit data
254
map to antennas, frequency channels
255
T/R comp bank
256
T/R comp alg
257
inverse freq chan bank
modulated transmit data
258
T/R packet trigger
GPS time transfer

Fig. 8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+/-
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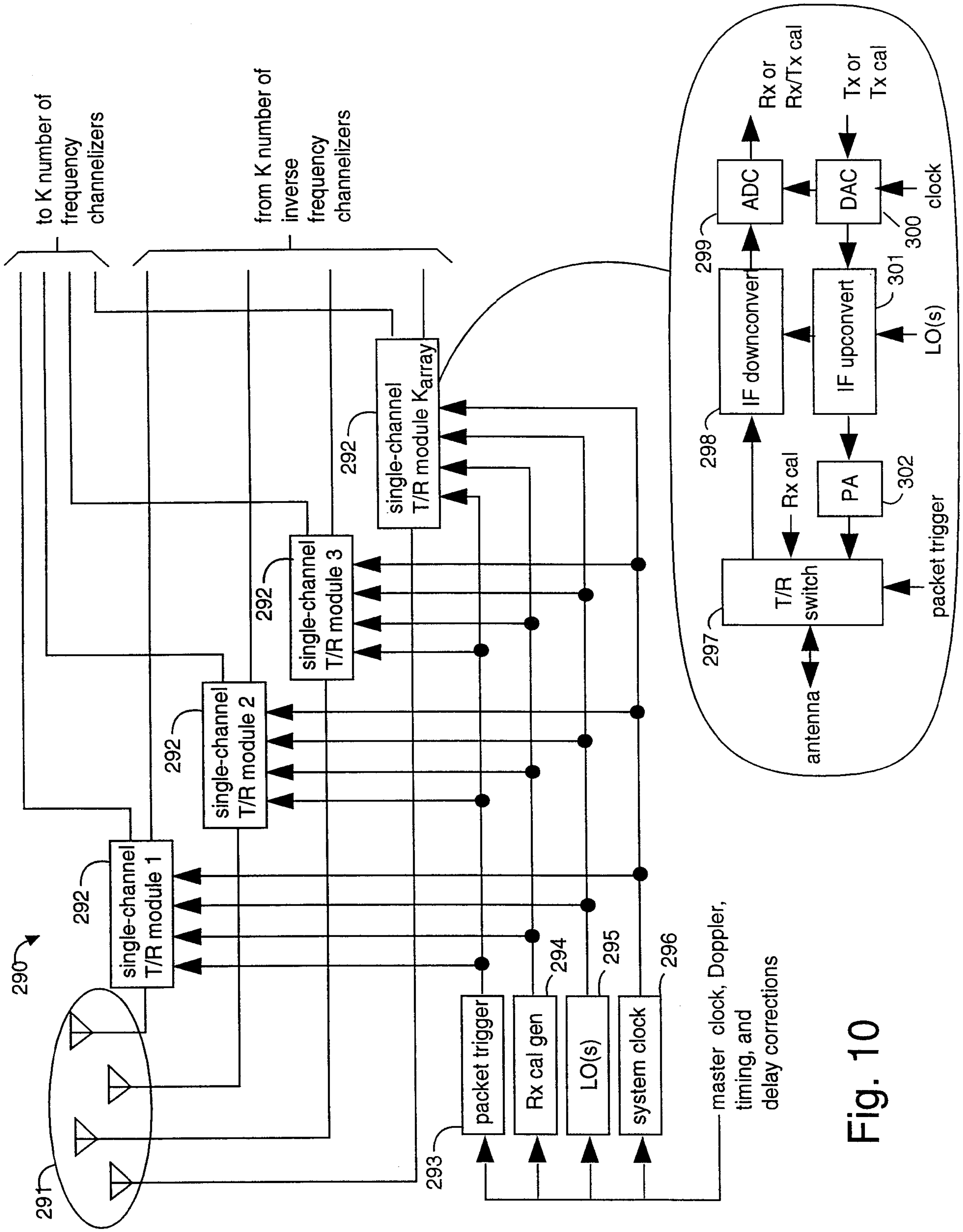
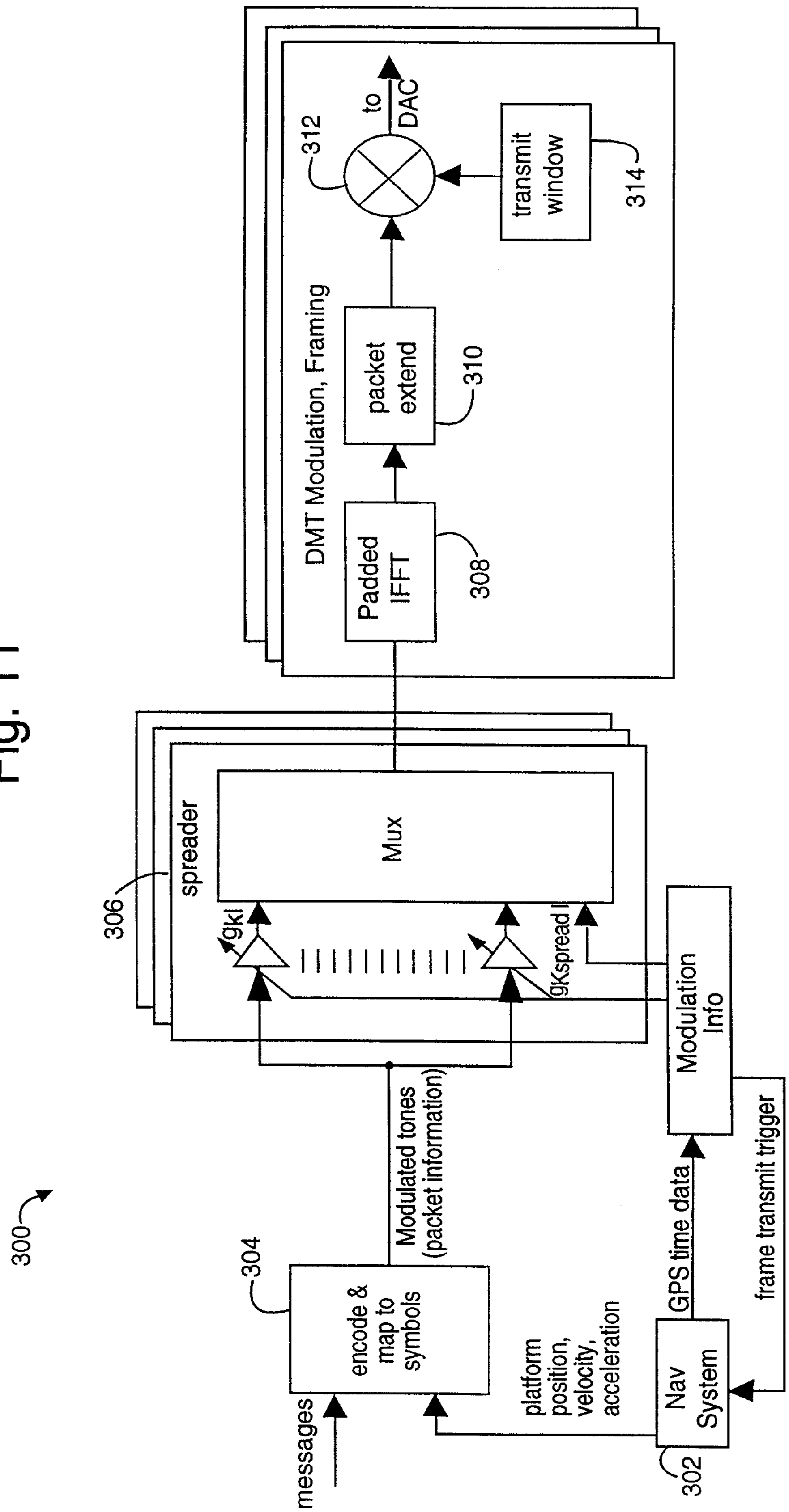
T/R front end
266
freq chan bank
268
map to data cells
demodulated receive data
270
weight adaptation algorithm
272
(multiantenna) despreader W
despread link message
273
delay, Doppler estimator
274
delay, Doppler equalize
276
symbol decode
decoded link message
278
symbol encode
link message
280
delay, Doppler preemphasis
282
(multiantenna) spreader G
spread transmit data
284
map to antennas, frequency channels
285
T/R comp bank
286
T/R comp alg
287
inverse freq chan bank
modulated transmit data
288
T/R packet trigger
GPS time transfer

Fig. 9



to K number of frequency channelizers
from K number of inverse frequency channelizers
single-channel T/R module 1
single-channel T/R module 2
single-channel T/R module 3
single-channel T/R module Karray
292
290
291
packet trigger
Rx cal gen
LO(s)
system clock
293
294
295
296
master clock, Doppler, timing, and delay corrections
antenna
T/R switch
297
packet trigger
Rx cal
PA
302
IF downconvert
298
IF upconvert
301
LO(s)
ADC
299
DAC
300
clock
Rx or Rx/Tx cal
Tx or Tx cal

Fig. 10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messages
encode & map to symbols
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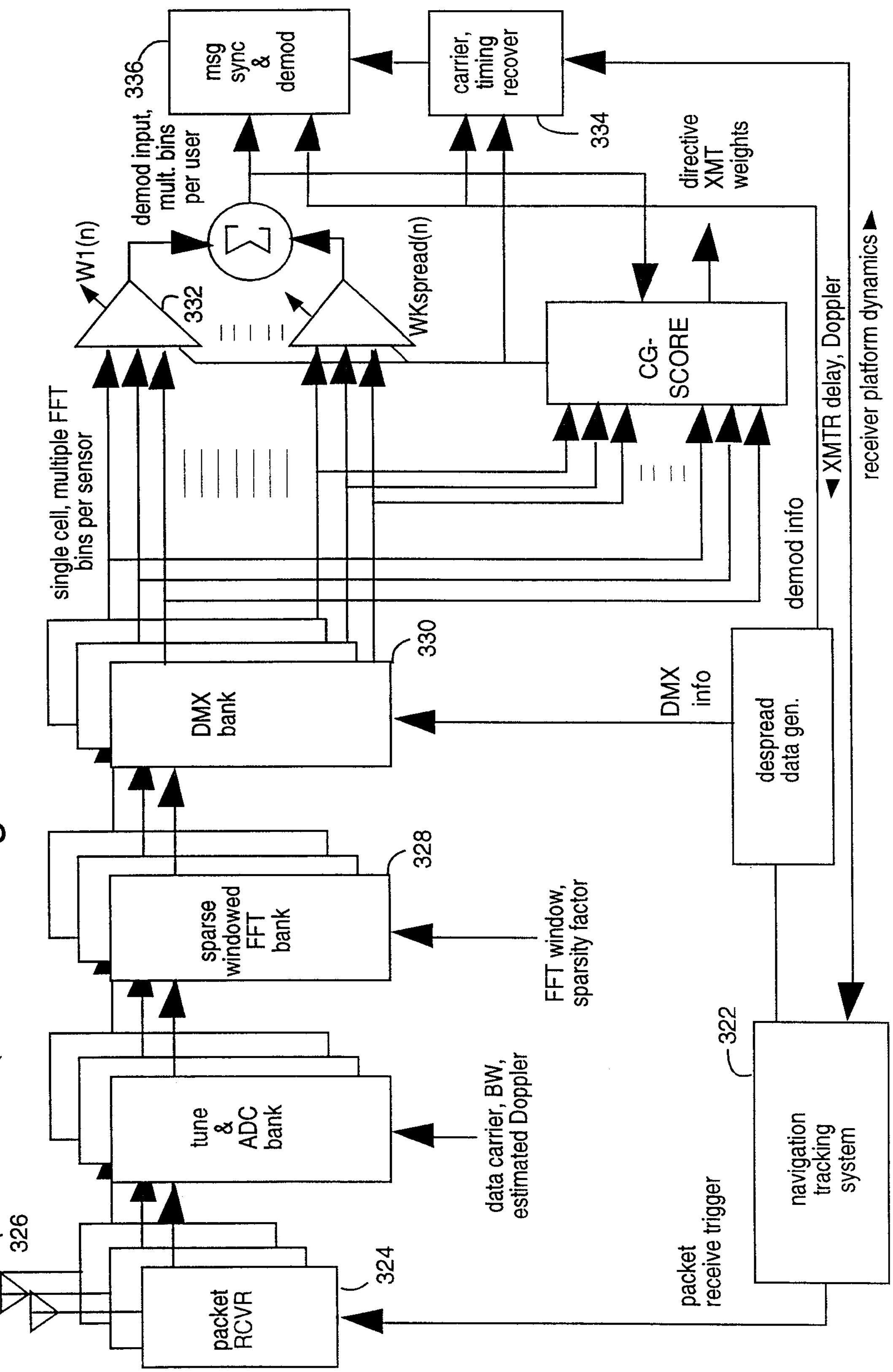
platform position, velocity, acceleration
Nav System
302
GPS time data
frame transmit trigger
Modulation Info
Modulated tones (packet information)
spreader
306
gkl
gKspread l
Mux
DMT Modulation, Framing
Padded IFFT
308
packet extend
310
312
to DAC
transmit window
314

Fig. 11

Fig. 12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sparse windowed FFT bank
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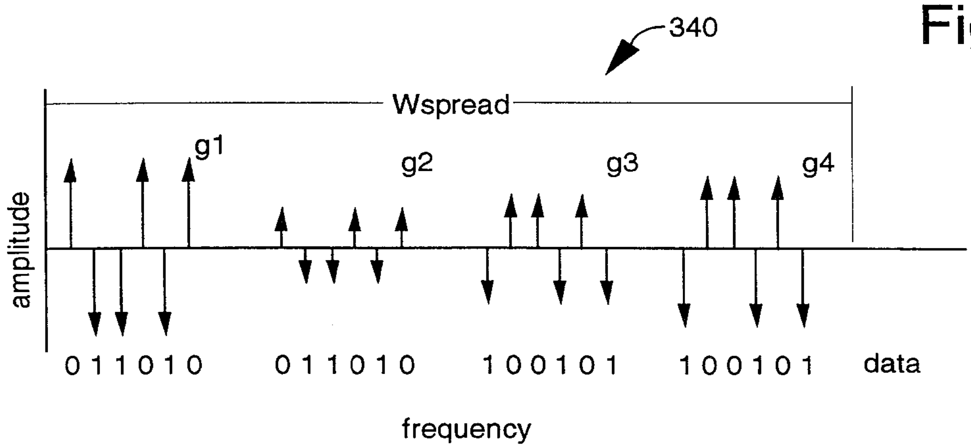
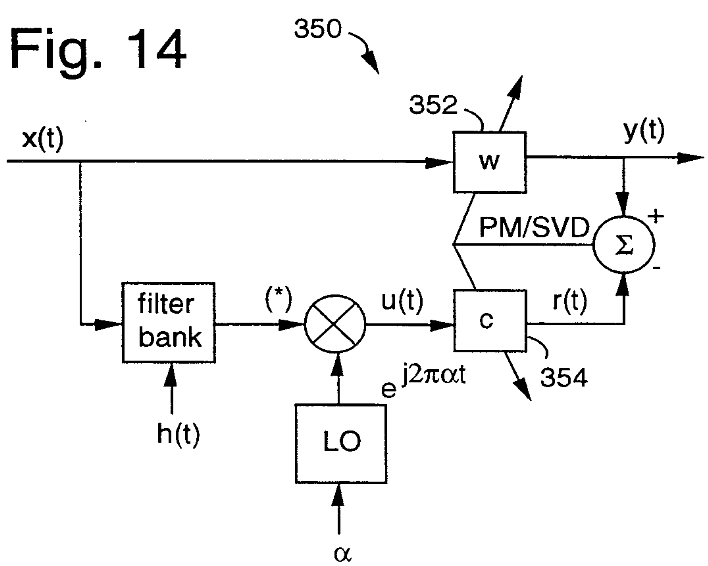
FFT window, sparsity factor
DMX bank
330
DMX info
despread data gen.
single cell, multiple FFT bins per sensor
demod info
W1(n)
332
WKspread(n)
CG-SCORE
demod input, mult. bins per user
msg sync & demod
336
carrier, timing recover
334
directive XMT weights
XMTR delay, Doppler
receiver platform dynamics
navigation tracking system
322
packet receive trigger



340
Wspread
g1
g2
g3
g4
amplitude
0 1 1 0 1 0
0 1 1 0 1 0
1 0 0 1 0 1
1 0 0 1 0 1
data
frequency

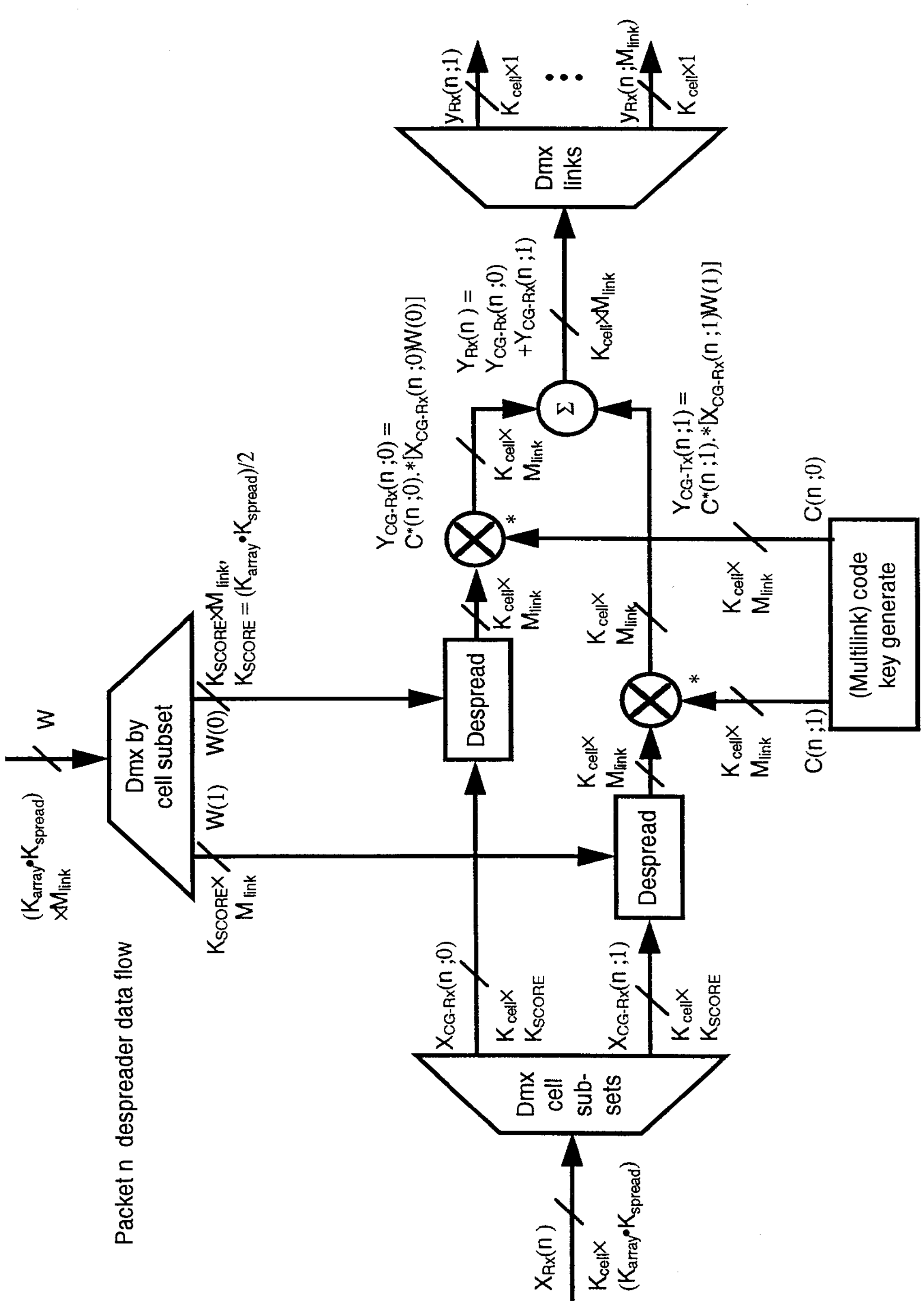
Fig. 13

Fig. 14



350
352
x(t)
w
y(t)
PM/SVD
Σ
+
-
filter bank
(*)
u(t)
c
r(t)
354
h(t)
e j2παt
LO
α

Fig. 15



Packet n despreader data flow
W
(Karray•Kspread) ×Mlink
Dmx by cell subset
W(0)
W(1)
KSCORE×Mlink, KSCORE = (Karray•Kspread)/2
KSCORE× Mlink
Despread
Despread
XRx(n)
Kcell× (Karray•Kspread)
Dmx cell sub-sets
XCG-Rx(n;0)
Kcell× KSCORE
XCG-Rx(n;1)
Kcell× KSCORE
Kcell× Mlink
YCG-Rx(n;0) = C*(n;0).*[XCG-Rx(n;0)W(0)]
YCG-Tx(n;1) = C*(n;1).*[XCG-Rx(n;1)W(1)]
C(n;0)
C(n;1)
(Multilink) code key generate
Σ
YRx(n) = YCG-Rx(n;0) +YCG-Rx(n;1)
Kcell×Mlink
Dmx links
yRx(n;1)
Kcell×1
yRx(n;Mlink)
Kcell×1

Fig. 16

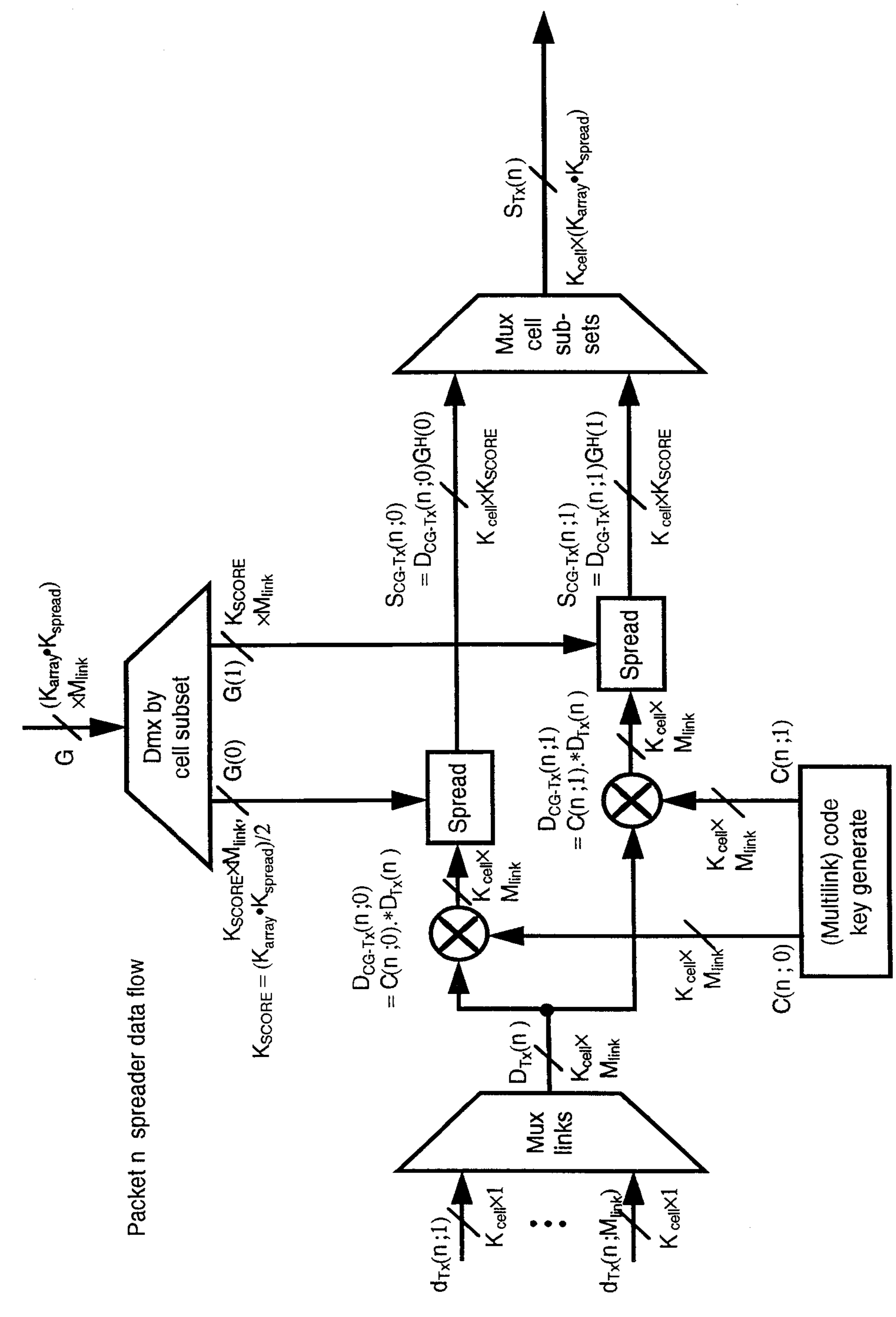

Packet n spreader data flow
$d_{Tx}(n;1)$
$K_{cell} \times 1$
$d_{Tx}(n;M_{link})$
$K_{cell} \times 1$
Mux links
$D_{Tx}(n)$
$K_{cell} \times M_{link}$
$D_{CG\text{-}Tx}(n;0) = C(n;0).*D_{Tx}(n)$
$K_{cell} \times M_{link}$
$D_{CG\text{-}Tx}(n;1) = C(n;1).*D_{Tx}(n)$
$K_{cell} \times M_{link}$
$K_{cell} \times M_{link}$
$K_{cell} \times M_{link}$
$C(n;0)$
$C(n;1)$
(Multilink) code key generate
G
$(K_{array} \bullet K_{spread}) \times M_{link}$
Dmx by cell subset
G(0)
G(1)
$K_{SCORE} \times M_{link}$,
$K_{SCORE} = (K_{array} \bullet K_{spread})/2$
$K_{SCORE} \times M_{link}$
Spread
Spread
$S_{CG\text{-}Tx}(n;0) = D_{CG\text{-}Tx}(n;0)G^H(0)$
$K_{cell} \times K_{SCORE}$
$S_{CG\text{-}Tx}(n;1) = D_{CG\text{-}Tx}(n;1)G^H(1)$
$K_{cell} \times K_{SCORE}$
Mux cell sub-sets
$S_{Tx}(n)$
$K_{cell} \times (K_{array} \bullet K_{spread})$

Fig. 17

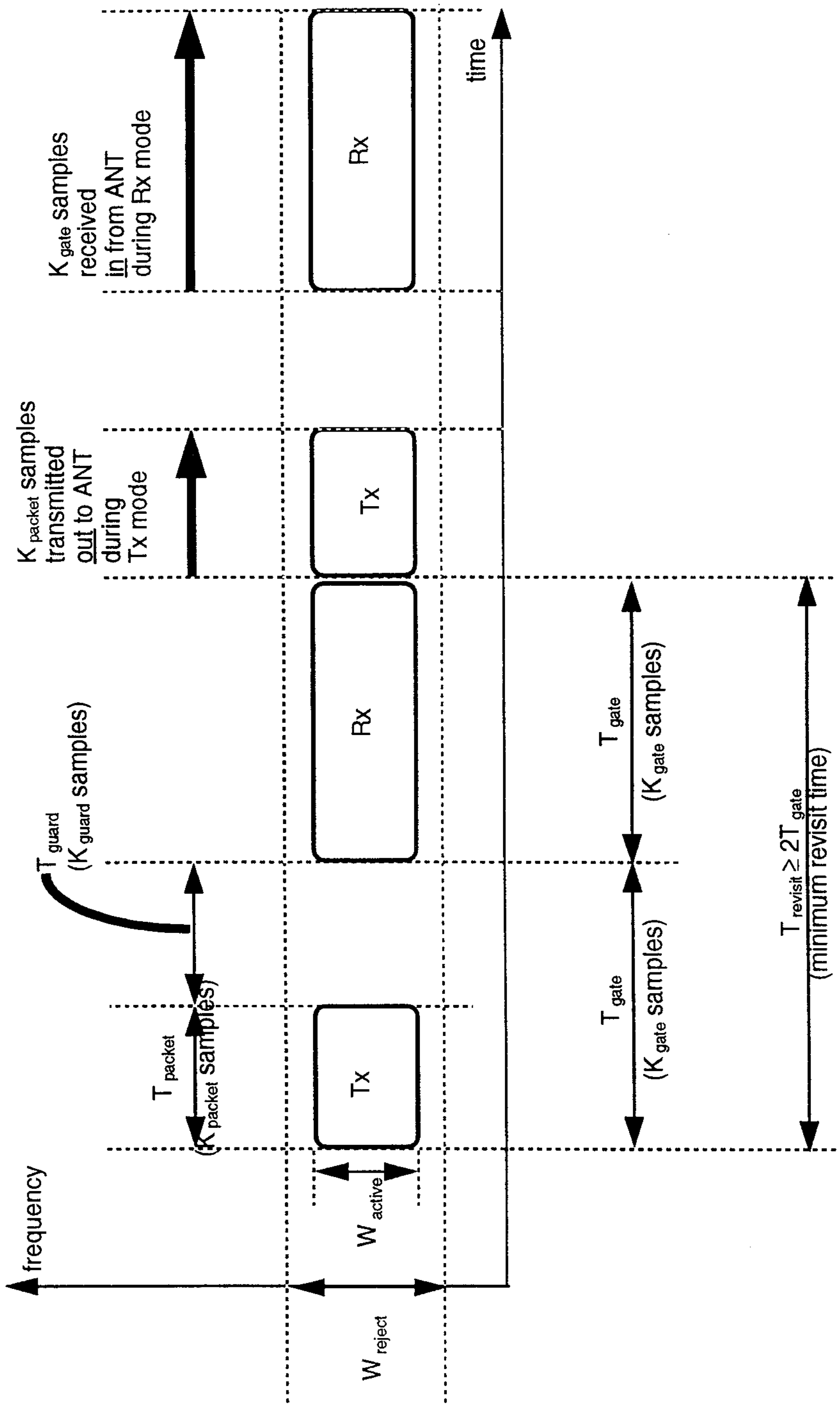

frequency
time
W_{reject}
W_{active}
Tx
Rx
Tx
Rx
T_{packet} (K_{packet} samples)
T_{guard} (K_{guard} samples)
K_{packet} samples transmitted out to ANT during Tx mode
K_{gate} samples received in from ANT during Rx mode
T_{gate} (K_{gate} samples)
T_{gate} (K_{gate} samples)
$T_{revisit} \geq 2T_{gate}$ (minimum revisit time)

Fig. 18

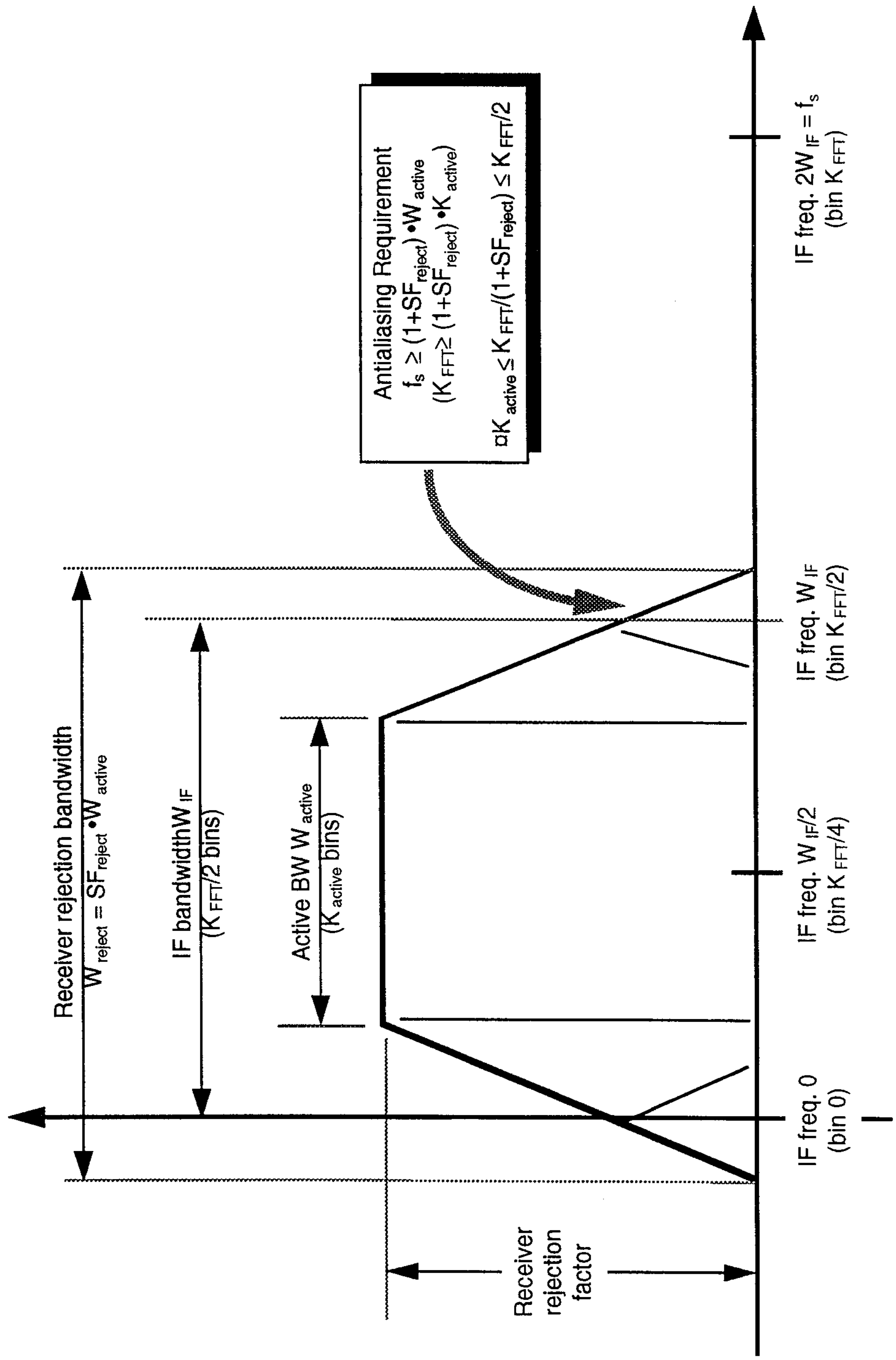

Antialiasing Requirement
$f_s \geq (1+SF_{reject}) \bullet W_{active}$
$(K_{FFT} \geq (1+SF_{reject}) \bullet K_{active})$
$K_{active} \leq K_{FFT}/(1+SF_{reject}) \leq K_{FFT}/2$
Receiver rejection bandwidth
$W_{reject} = SF_{reject} \bullet W_{active}$
IF bandwidth W_{IF}
($K_{FFT}/2$ bins)
Active BW W_{active}
(K_{active} bins)
Receiver rejection factor
IF freq. 0
(bin 0)
IF freq. $W_{IF}/2$
(bin $K_{FFT}/4$)
IF freq. W_{IF}
(bin $K_{FFT}/2$)
IF freq. $2W_{IF} = f_s$
(bin K_{FFT})

Fig. 19

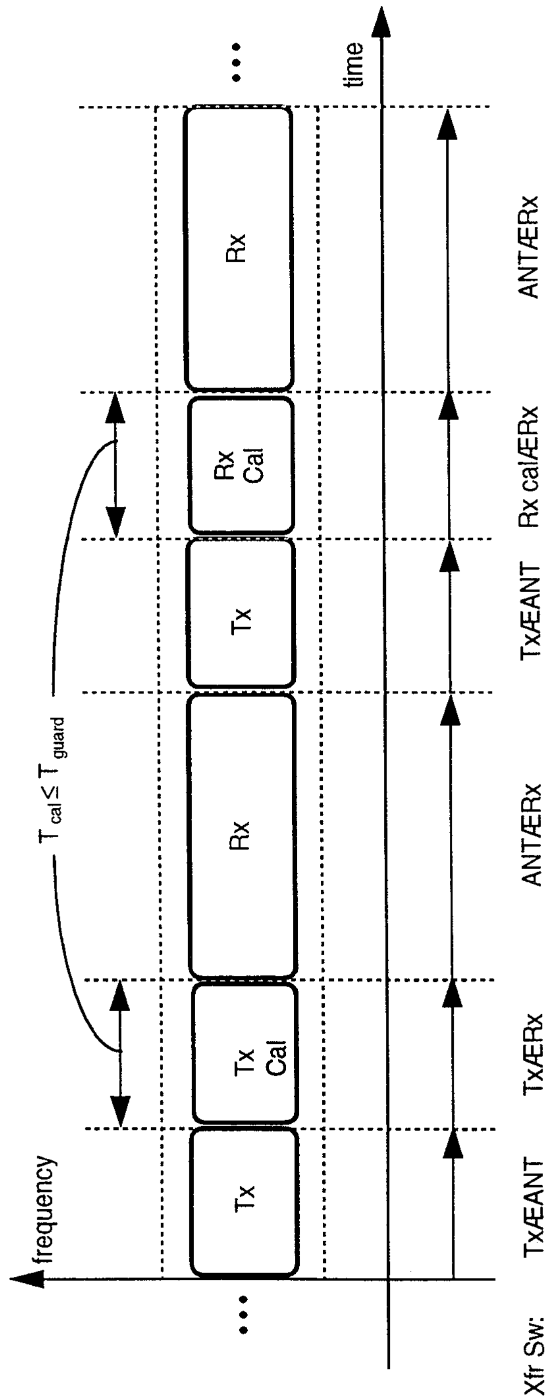
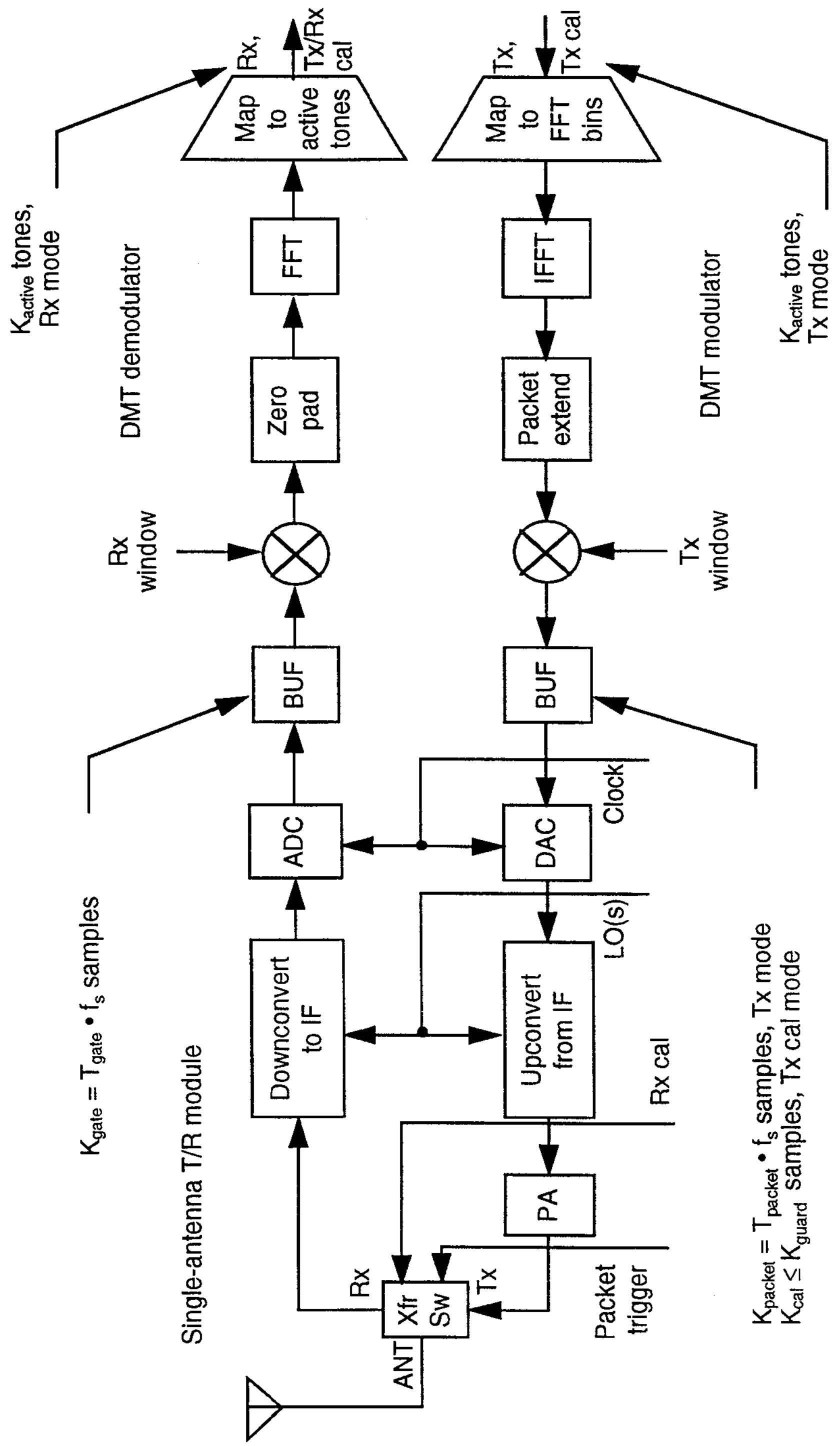

frequency
time
Tx
Tx
Cal
Rx
Tx
Rx
Cal
Rx
$T_{cal} \le T_{guard}$
Xfr Sw:
TxÆANT
TxÆRx
ANTÆRx
TxÆANT
Rx calÆRx
ANTÆRx

Fig. 20


Map to active tones
Rx, Tx/Rx cal
FFT
Zero pad
Rx window
BUF
ADC
Downconvert to IF
Xfr Sw
Rx
Tx
ANT
Map to FFT bins
Tx, Tx cal
IFFT
Packet extend
Tx window
BUF
DAC
Clock
Upconvert from IF
LO(s)
PA
Rx cal
Packet trigger
K_{active} tones, Rx mode
DMT demodulator
$K_{gate} = T_{gate} \bullet f_s$ samples
Single-antenna T/R module
DMT modulator
K_{active} tones, Tx mode
$K_{packet} = T_{packet} \bullet f_s$ samples, Tx mode
$K_{cal} \le K_{guard}$ samples, Tx cal mode

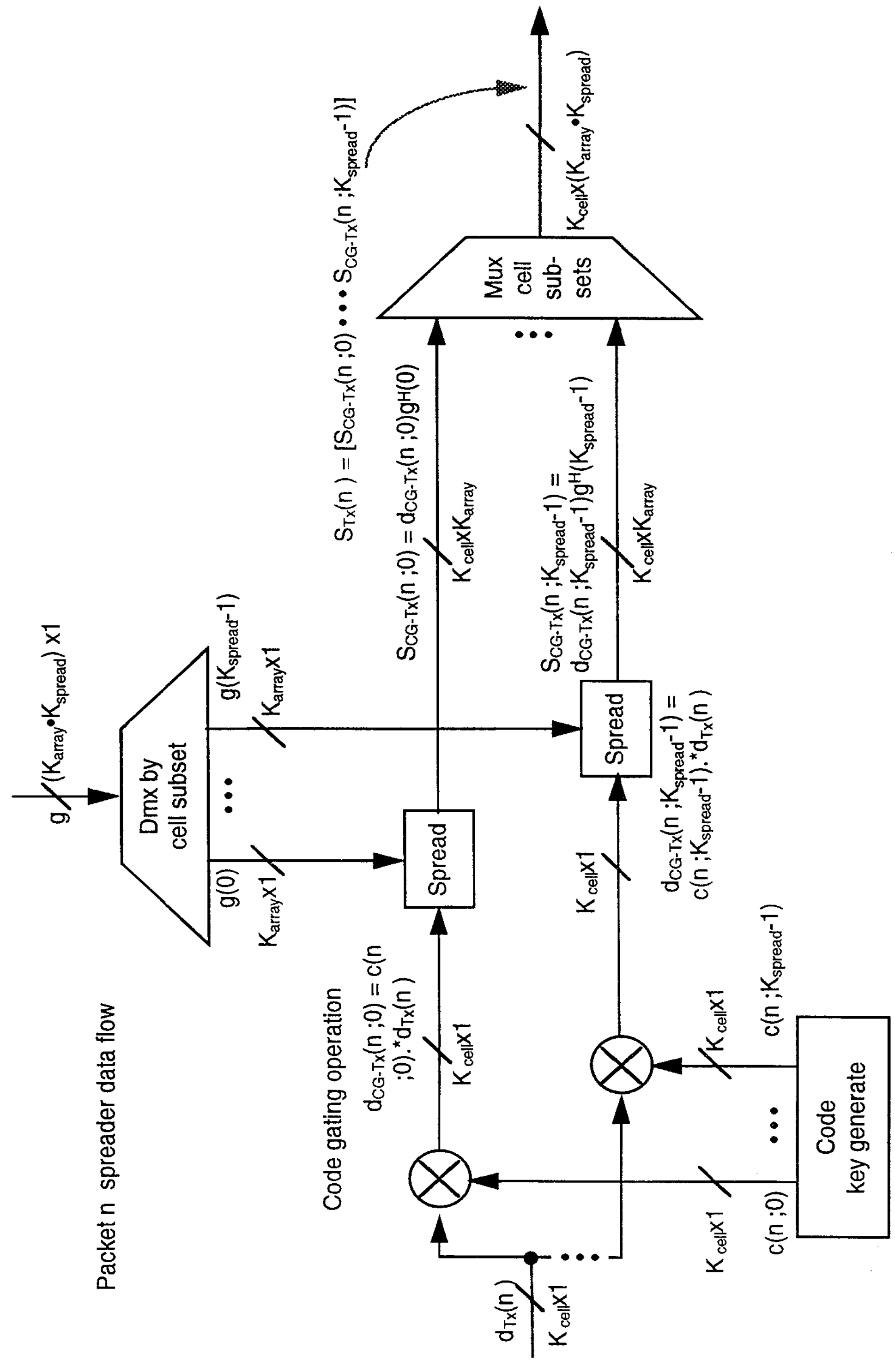

Packet n spreader data flow
Code gating operation
$d_{CG-Tx}(n;0) = c(n;0).*d_{Tx}(n)$
$d_{Tx}(n)$
$K_{cell}X1$
g
$(K_{array} \bullet K_{spread})\ X1$
Dmx by cell subset
$g(0)$
$K_{array}X1$
$g(K_{spread}-1)$
Spread
Spread
$d_{CG-Tx}(n;K_{spread}-1) = c(n;K_{spread}-1).*d_{Tx}(n)$
$c(n;0)$
$c(n;K_{spread}-1)$
Code key generate
$S_{CG-Tx}(n;0) = d_{CG-Tx}(n;0)g^H(0)$
$S_{CG-Tx}(n;K_{spread}-1) = d_{CG-Tx}(n;K_{spread}-1)g^H(K_{spread}-1)$
$K_{cell}XK_{array}$
Mux cell sub-sets
$S_{Tx}(n) = [S_{CG-Tx}(n;0) \bullet\bullet\bullet S_{CG-Tx}(n;K_{spread}-1)]$
$K_{cell}X(K_{array} \bullet K_{spread})$

Fig. 21

Fig. 22

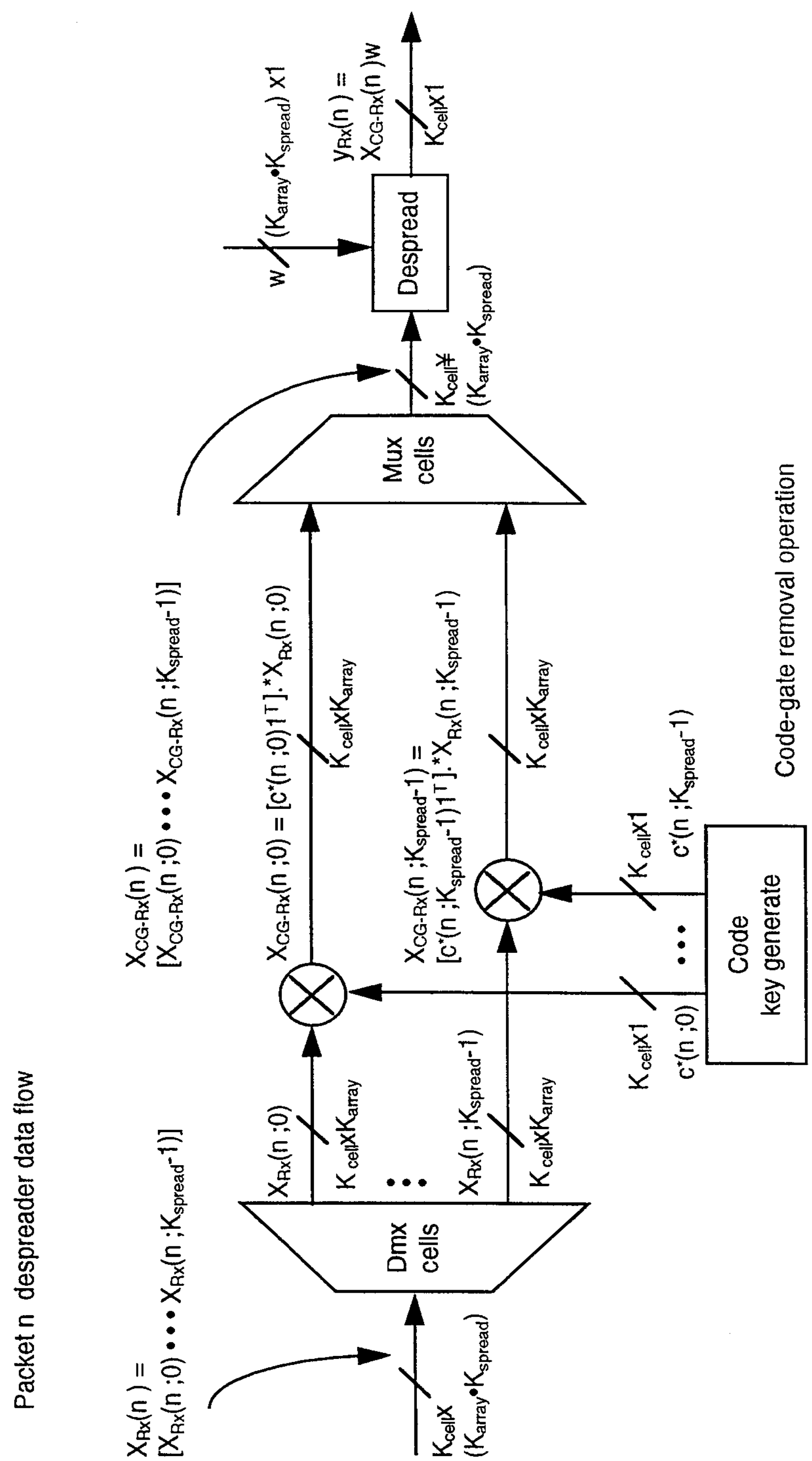

Packet n despreader data flow
$X_{Rx}(n) = [X_{Rx}(n;0) \bullet\bullet\bullet X_{Rx}(n;K_{spread}-1)]$
$K_{cell}X(K_{array}\bullet K_{spread})$
Dmx cells
$X_{Rx}(n;0)$
$K_{cell}XK_{array}$
$X_{Rx}(n;K_{spread}-1)$
$K_{cell}XK_{array}$
$X_{CG\text{-}Rx}(n) = [X_{CG\text{-}Rx}(n;0) \bullet\bullet\bullet X_{CG\text{-}Rx}(n;K_{spread}-1)]$
$X_{CG\text{-}Rx}(n;0) = [c^*(n;0)1^T].*X_{Rx}(n;0)$
$X_{CG\text{-}Rx}(n;K_{spread}-1) = [c^*(n;K_{spread}-1)1^T].*X_{Rx}(n;K_{spread}-1)$
$K_{cell}XK_{array}$
$K_{cell}XK_{array}$
$K_{cell}X1$
$c^*(n;0)$
$K_{cell}X1$
$c^*(n;K_{spread}-1)$
Code key generate
Code-gate removal operation
Mux cells
$K_{cell}X(K_{array}\bullet K_{spread})$
Despread
w
$(K_{array}\bullet K_{spread})\ X1$
$y_{Rx}(n) = X_{CG\text{-}Rx}(n)w$
$K_{cell}X1$

Fig. 23

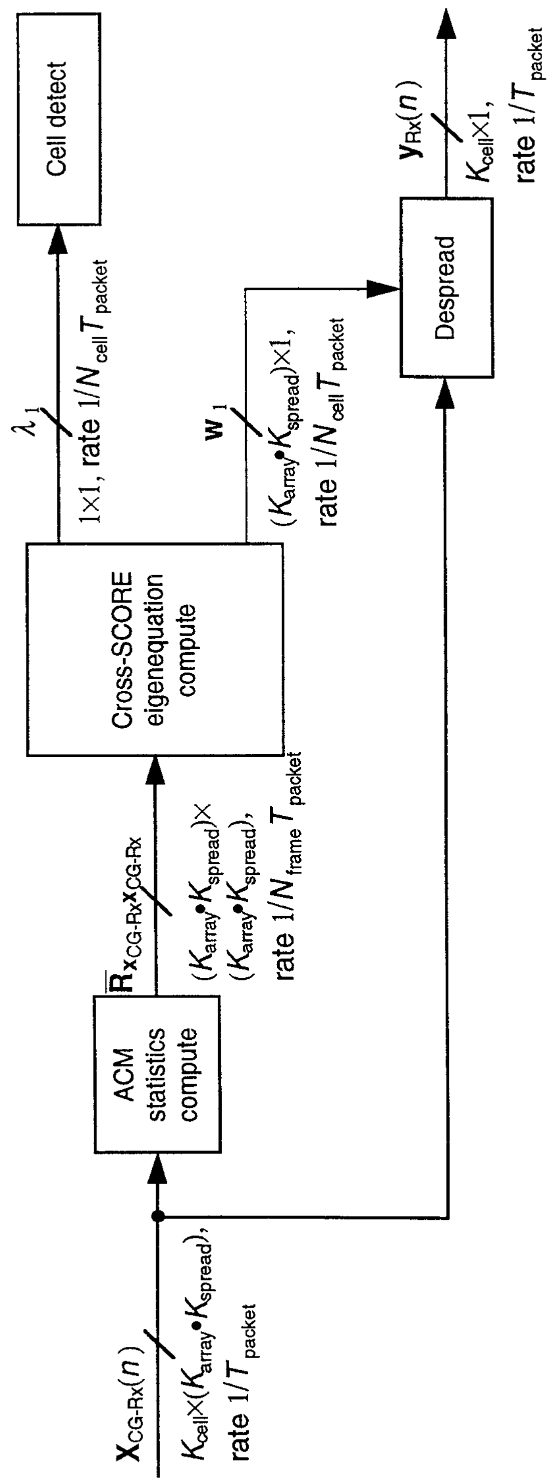
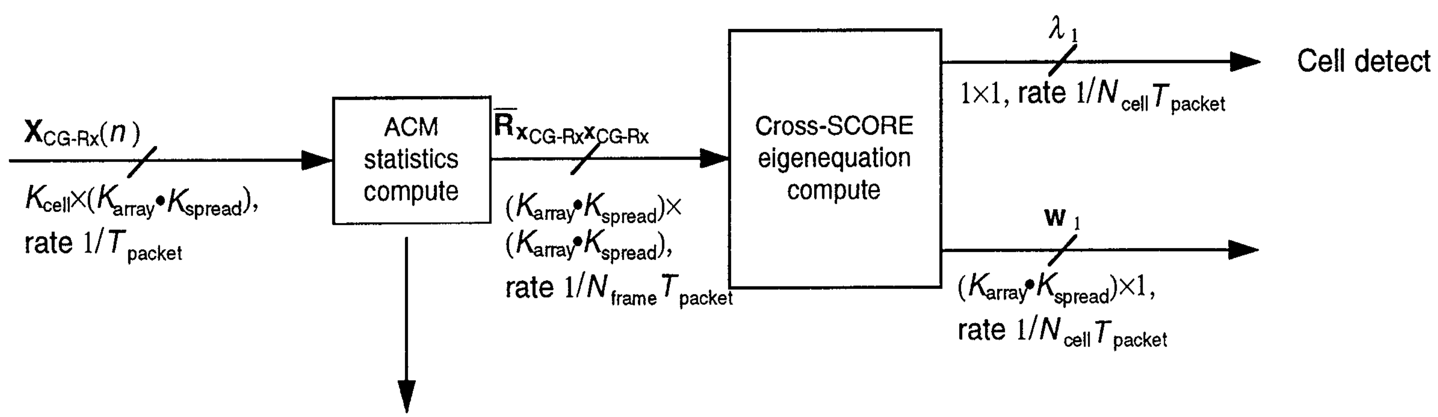

Cell detect
λ_1
1×1, rate $1/N_{cell}T_{packet}$
Despread
$\mathbf{y}_{Rx}(n)$
$K_{cell}\times 1$,
rate $1/T_{packet}$
$\mathbf{w}_1$
$(K_{array}\bullet K_{spread})\times 1$,
rate $1/N_{cell}T_{packet}$
Cross-SCORE eigenequation compute
$\overline{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}$
$(K_{array}\bullet K_{spread})\times$
$(K_{array}\bullet K_{spread})$,
rate $1/N_{frame}T_{packet}$
ACM statistics compute
$\mathbf{x}_{CG\text{-}Rx}(n)$
$K_{cell}\times(K_{array}\bullet K_{spread})$,
rate $1/T_{packet}$

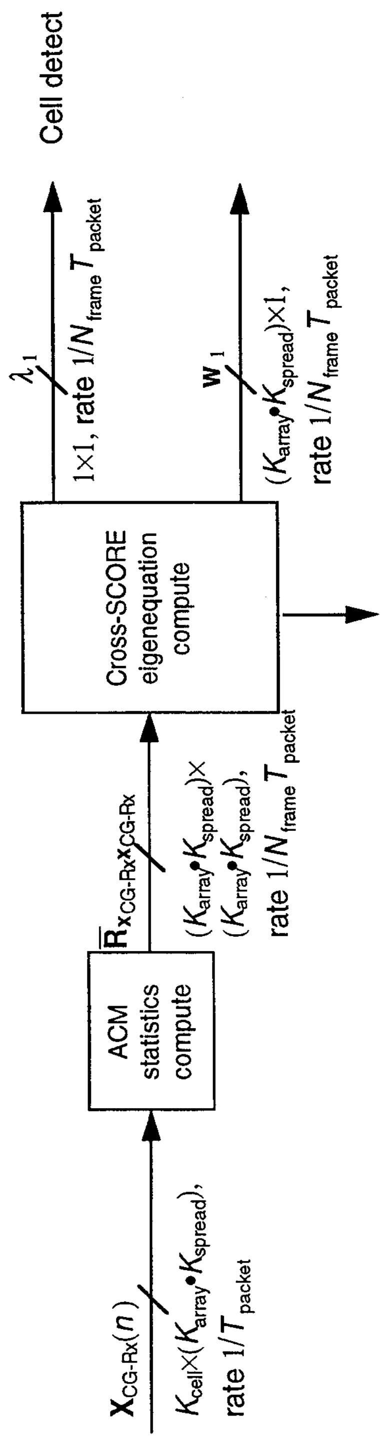
Fig. 24



$$\overline{\mathbf{R}}_{\mathbf{x}_{\text{CG-Rx}}\mathbf{x}_{\text{CG-Rx}}}(k,l) = \frac{1}{K_{\text{cell}}N_{\text{frame}}}\sum_{q=0}^{N_{\text{frame}}-1}\mathbf{X}^{H}_{\text{CG-Rx}}(N_{\text{frame}}n+q\,;k)\,\mathbf{X}_{\text{CG-Rx}}(N_{\text{frame}}n+q\,;l)$$

$$\overline{\mathbf{R}}_{\mathbf{x}_{\text{CG-Rx}}\mathbf{x}_{\text{CG-Rx}}} = \begin{bmatrix} \overline{\mathbf{R}}_{\mathbf{x}_{\text{CG-Rx}}\mathbf{x}_{\text{CG-Rx}}}(0,0) & \cdots & \overline{\mathbf{R}}_{\mathbf{x}_{\text{CG-Rx}}\mathbf{x}_{\text{CG-Rx}}}(0,K_{\text{spread}}-1) \\ \vdots & \ddots & \vdots \\ \overline{\mathbf{R}}_{\mathbf{x}_{\text{CG-Rx}}\mathbf{x}_{\text{CG-Rx}}}(K_{\text{spread}}-1,0) & \cdots & \overline{\mathbf{R}}_{\mathbf{x}_{\text{CG-Rx}}\mathbf{x}_{\text{CG-Rx}}}(K_{\text{spread}}-1,K_{\text{spread}}-1) \end{bmatrix}$$

Fig. 25



$$\widetilde{\mathbf{R}}_{\mathbf{x}_{\text{CG-Rx}}\mathbf{x}_{\text{CG-Rx}}} = \operatorname{diag}\left\{\overline{\mathbf{R}}_{\mathbf{x}_{\text{CG-Rx}}\mathbf{x}_{\text{CG-Rx}}}(k,k)\right\}$$

$$= \begin{bmatrix} \overline{\mathbf{R}}_{\mathbf{x}_{\text{CG-Rx}}\mathbf{x}_{\text{CG-Rx}}}(0,0) & & 0 \\ & \ddots & \\ 0 & & \overline{\mathbf{R}}_{\mathbf{x}_{\text{CG-Rx}}\mathbf{x}_{\text{CG-Rx}}}(K_{\text{spread}}-1, K_{\text{spread}}-1) \end{bmatrix}$$

$$\lambda_q \widetilde{\mathbf{R}}_{\mathbf{x}_{\text{CG-Rx}}\mathbf{x}_{\text{CG-Rx}}} \mathbf{w}_q = \overline{\mathbf{R}}_{\mathbf{x}_{\text{CG-Rx}}\mathbf{x}_{\text{CG-Rx}}} \mathbf{w}_q, \quad \lambda_q \geq \lambda_{q+1}$$

$$\mathbf{w}_q^H \overline{\mathbf{R}}_{\mathbf{x}_{\text{CG-Rx}}\mathbf{x}_{\text{CG-Rx}}} \mathbf{w}_q = 1 \quad \text{(unit output power constraint)}$$

Fig. 26

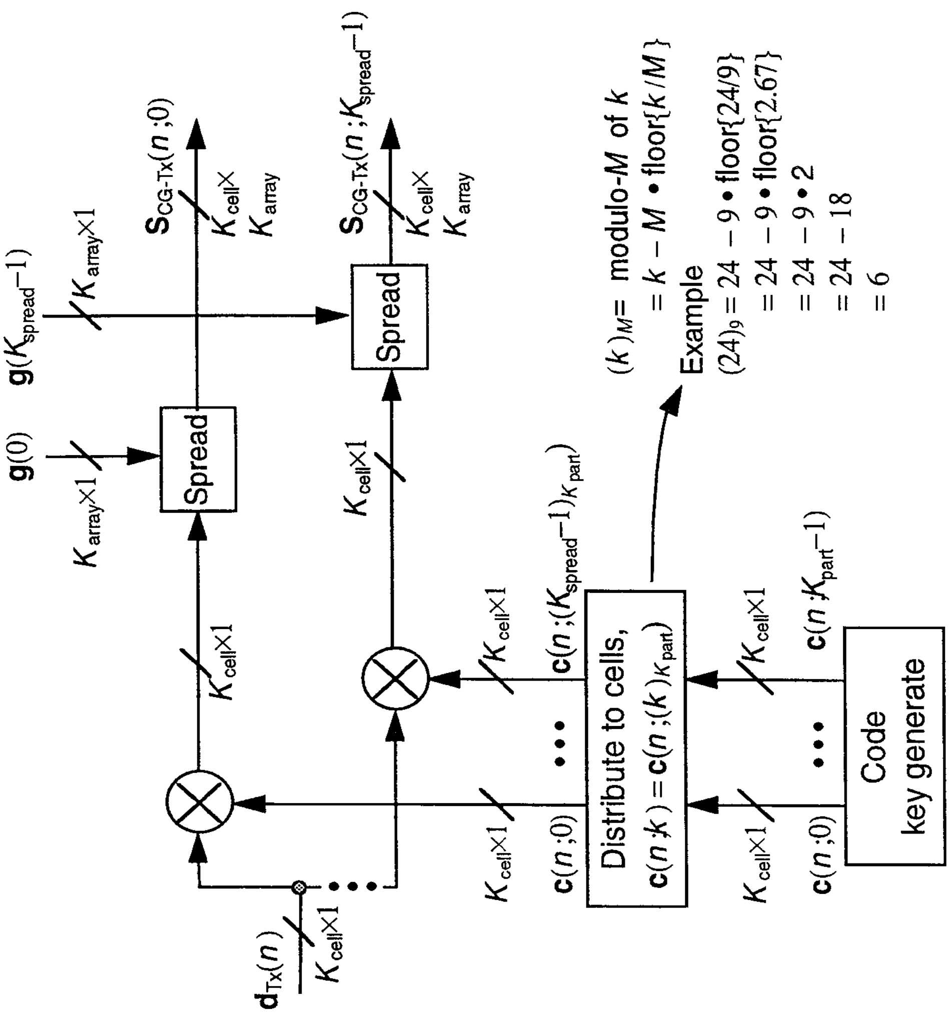
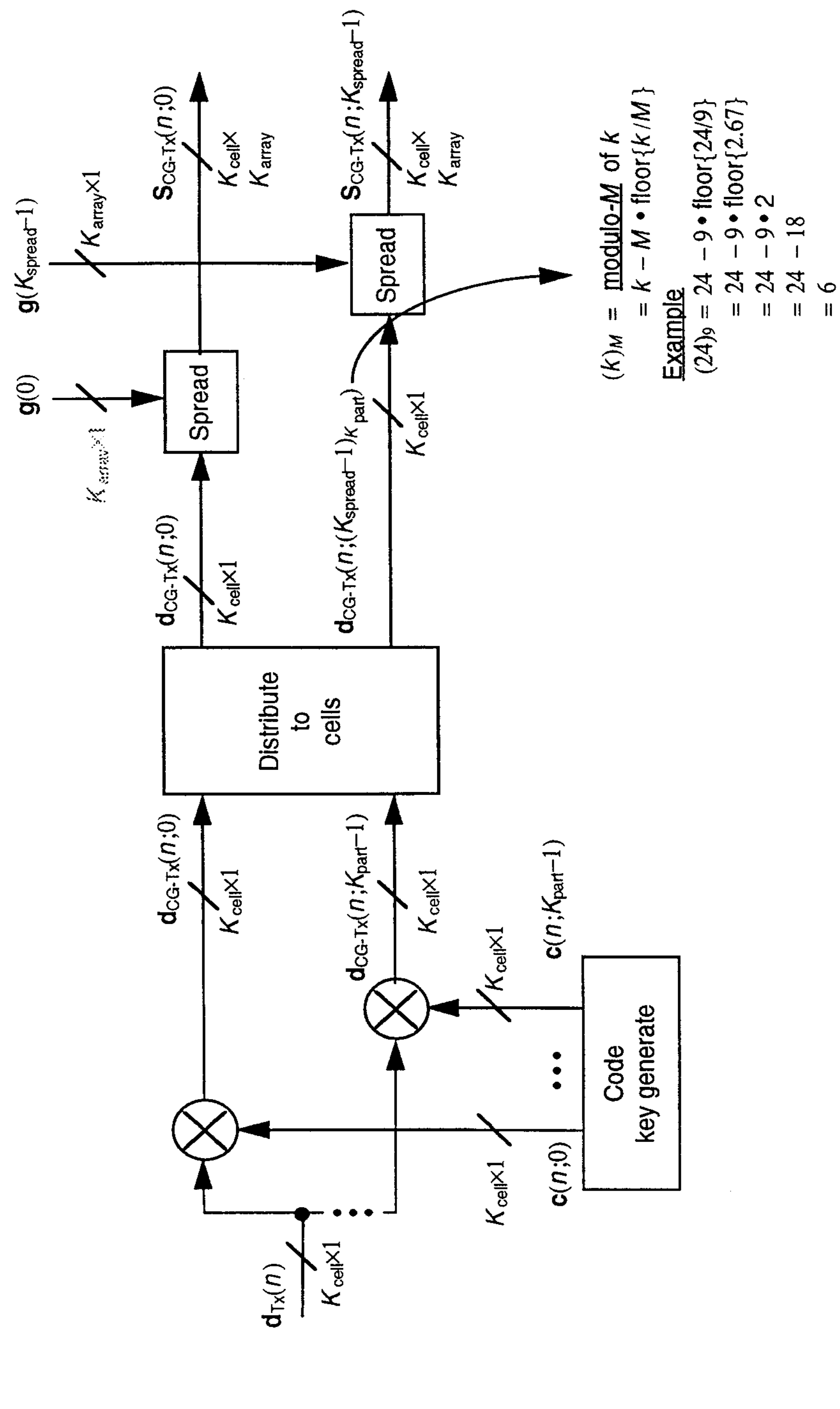
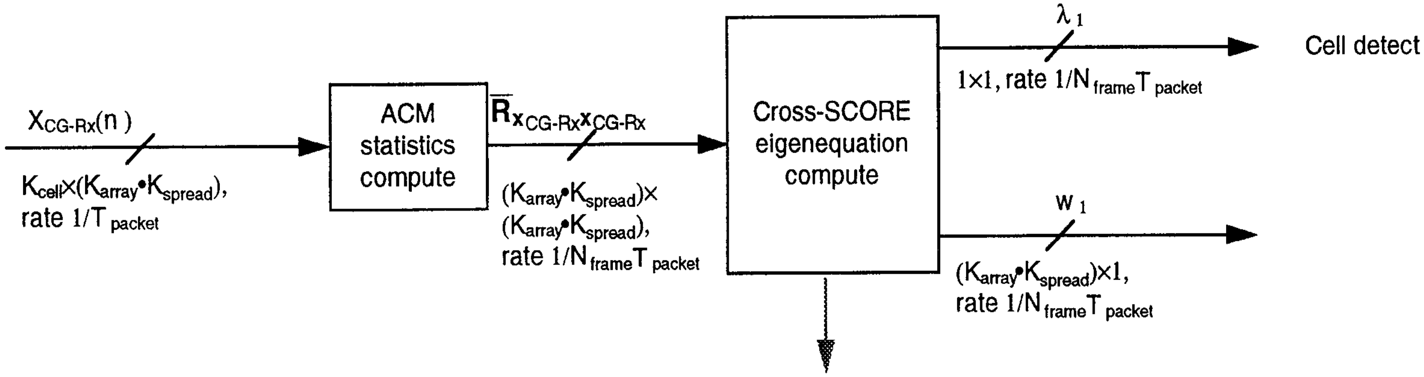

Spread
Spread
Distribute to cells,
Code
key generate
Example

Fig. 27



$\mathbf{d}_{Tx}(n)$
$K_{cell}\times 1$
Code key generate
$\mathbf{c}(n;0)$
$K_{cell}\times 1$
$\mathbf{c}(n;K_{part}-1)$
$K_{cell}\times 1$
$\mathbf{d}_{CG\text{-}Tx}(n;0)$
$K_{cell}\times 1$
$\mathbf{d}_{CG\text{-}Tx}(n;K_{part}-1)$
$K_{cell}\times 1$
Distribute to cells
$\mathbf{d}_{CG\text{-}Tx}(n;0)$
$K_{cell}\times 1$
$\mathbf{d}_{CG\text{-}Tx}(n;(K_{spread}-1)_{K_{part}})$
$K_{cell}\times 1$
$\mathbf{g}(0)$
$K_{array}\times 1$
$\mathbf{g}(K_{spread}-1)$
$K_{array}\times 1$
Spread
Spread
$\mathbf{s}_{CG\text{-}Tx}(n;0)$
$K_{cell}\times K_{array}$
$\mathbf{s}_{CG\text{-}Tx}(n;K_{spread}-1)$
$K_{cell}\times K_{array}$
$(k)_M$ = modulo-M of k
$= k - M \bullet \text{floor}\{k/M\}$
Example
$(24)_9 = 24 - 9 \bullet \text{floor}\{24/9\}$
$= 24 - 9 \bullet \text{floor}\{2.67\}$
$= 24 - 9 \bullet 2$
$= 24 - 18$
$= 6$

Fig. 28



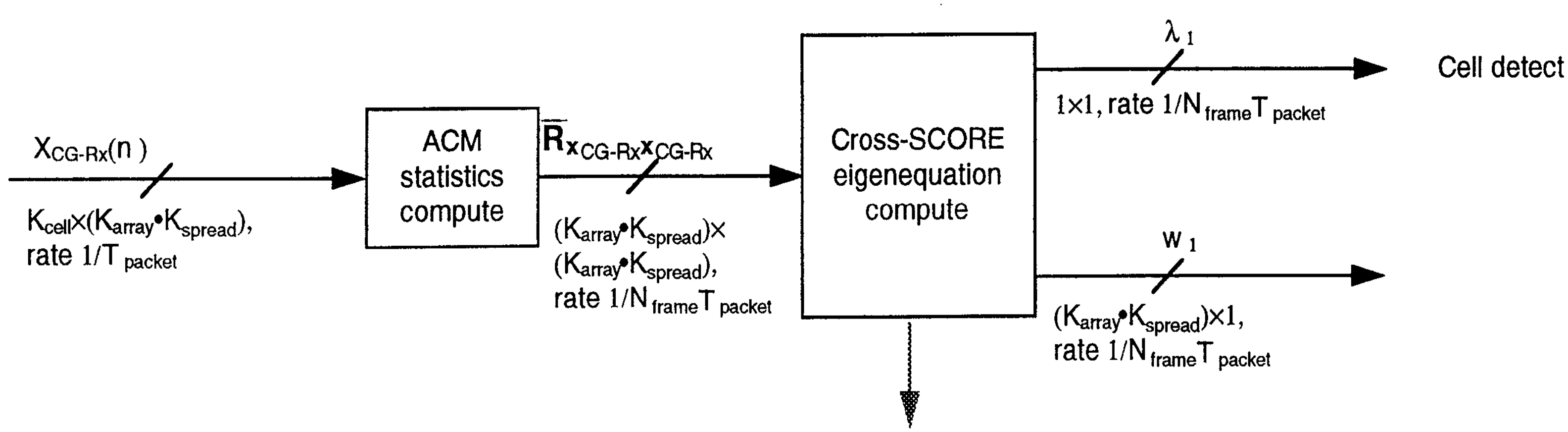
$$\widetilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}} = \begin{bmatrix} \widetilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(0,0) & \cdots & \widetilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(0,K_{spread}-1) \\ \vdots & \ddots & \vdots \\ \widetilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(K_{spread}-1,0) & \cdots & \widetilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(K_{spread}-1,K_{spread}-1) \end{bmatrix}$$

$$\widetilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(k,l) = \begin{cases} \widetilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(k,l), & k-l = \text{integer multiple of } K_{part} \\ 0, & \text{otherwise} \end{cases}$$

$$\lambda_q \widetilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}} \mathbf{w}_q = \overline{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}} \mathbf{w}_q, \quad \lambda_q \geq \lambda_{q+1}$$

$$\mathbf{w}_q^H \overline{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}} \mathbf{w}_q = 1 \quad \text{(unit output power constraint)}$$

Fig. 29



$$\tilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}} = \begin{bmatrix} \tilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(0,0) & \cdots & \tilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(0,K_{spread}-1) \\ \vdots & \ddots & \vdots \\ \tilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(K_{spread}-1,0) & \cdots & \tilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(K_{spread}-1,K_{spread}-1) \end{bmatrix}$$

$$\tilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(k,l) = \begin{cases} \tilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(k,l), & k-l = \text{integer multiple of } K_{part} \\ 0, & \text{otherwise} \end{cases}$$

$$\lambda_q \tilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}\mathbf{w}_q = \bar{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}\mathbf{w}_q,\ \lambda_q \geq \lambda_{q+1}$$

$$\mathbf{w}_q^H \bar{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}\mathbf{w}_q = 1 \quad \text{(unit output power constraint)}$$

Fig. 30

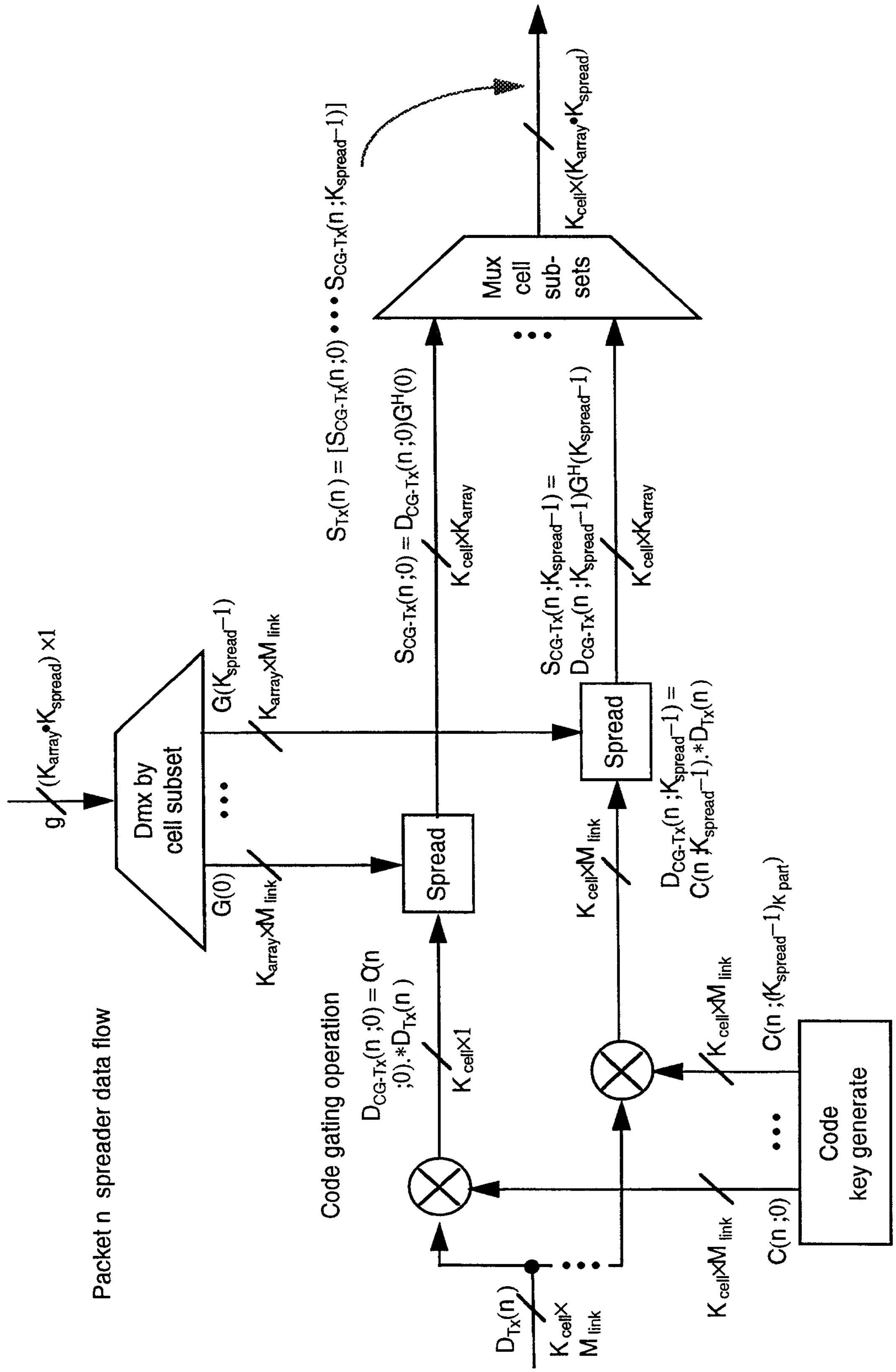

Packet n spreader data flow
Code gating operation
$D_{CG\text{-}Tx}(n;0) = C(n;0).*D_{Tx}(n)$
g
$(K_{array} \bullet K_{spread}) \times 1$
Dmx by cell subset
$G(0)$
$G(K_{spread}-1)$
$K_{array} \times M_{link}$
$K_{array} \times M_{link}$
Spread
Spread
$S_{CG\text{-}Tx}(n;0) = D_{CG\text{-}Tx}(n;0)G^H(0)$
$S_{CG\text{-}Tx}(n;K_{spread}-1) = D_{CG\text{-}Tx}(n;K_{spread}-1)G^H(K_{spread}-1)$
$D_{CG\text{-}Tx}(n;K_{spread}-1) = C(n;K_{spread}-1).*D_{Tx}(n)$
$S_{Tx}(n) = [S_{CG\text{-}Tx}(n;0) \bullet\bullet\bullet S_{CG\text{-}Tx}(n;K_{spread}-1)]$
$K_{cell} \times K_{array}$
$K_{cell} \times K_{array}$
Mux cell sub-sets
$K_{cell} \times (K_{array} \bullet K_{spread})$
$D_{Tx}(n)$
$K_{cell} \times M_{link}$
$K_{cell} \times 1$
$K_{cell} \times M_{link}$
$K_{cell} \times M_{link}$
$K_{cell} \times M_{link}$
$C(n;0)$
$C(n;(K_{spread}-1)_{K_{part}})$
Code key generate

Fig. 31

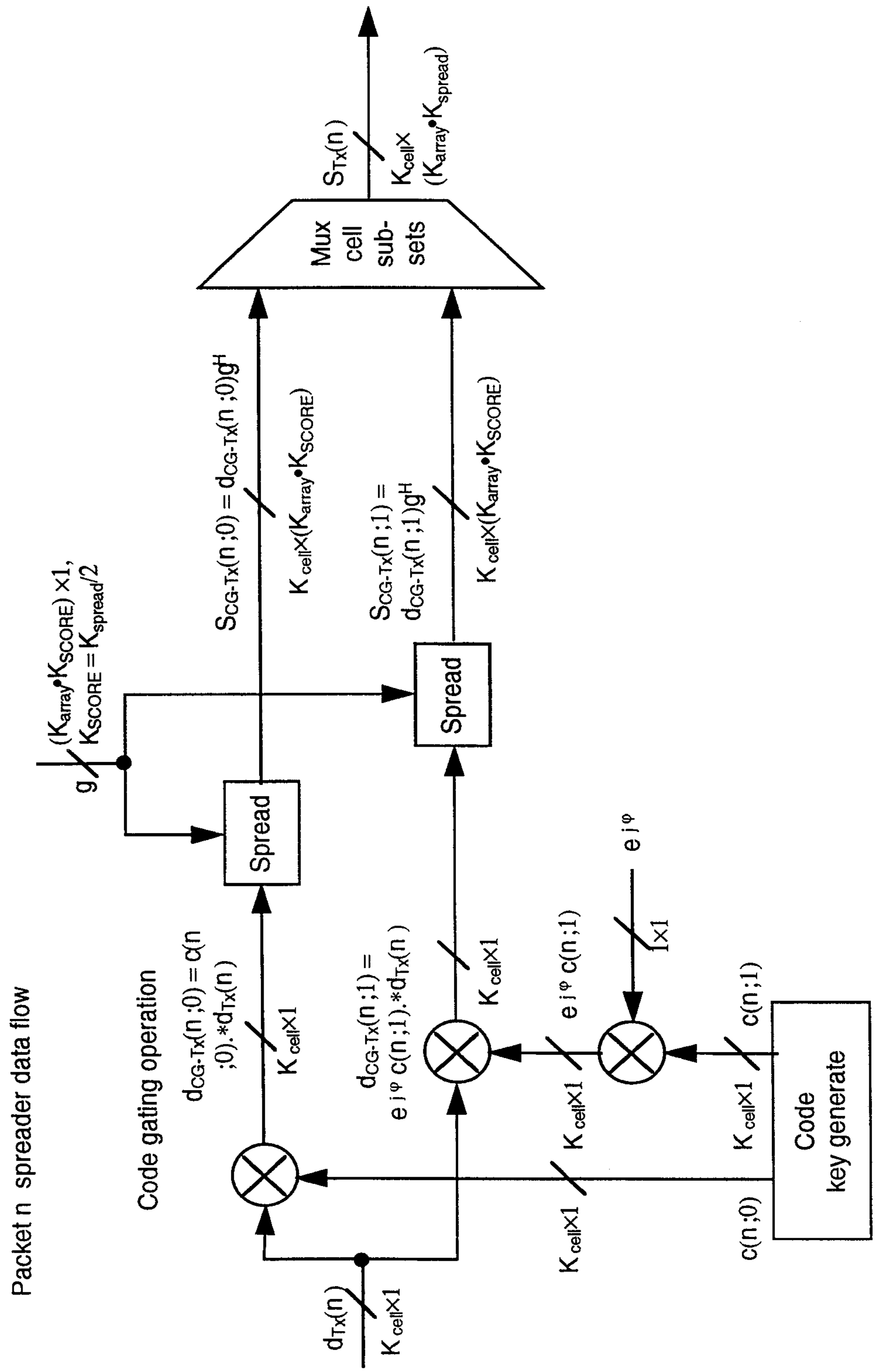

Packet n spreader data flow
Code gating operation
$d_{TX}(n)$
$K_{cell}\times1$
$d_{CG\text{-}TX}(n;0) = c(n;0).*d_{TX}(n)$
$d_{CG\text{-}TX}(n;1) = e^{j\varphi}c(n;1).*d_{TX}(n)$
$e^{j\varphi}$
1×1
$e^{j\varphi}c(n;1)$
c(n;0)
c(n;1)
Code key generate
g
$(K_{array}\bullet K_{SCORE})\times1,$
$K_{SCORE} = K_{spread}/2$
Spread
Spread
$S_{CG\text{-}TX}(n;0) = d_{CG\text{-}TX}(n;0)g^H$
$S_{CG\text{-}TX}(n;1) = d_{CG\text{-}TX}(n;1)g^H$
$K_{cell}\times(K_{array}\bullet K_{SCORE})$
Mux cell sub-sets
$S_{TX}(n)$
$K_{cell}\times(K_{array}\bullet K_{spread})$

Fig. 32

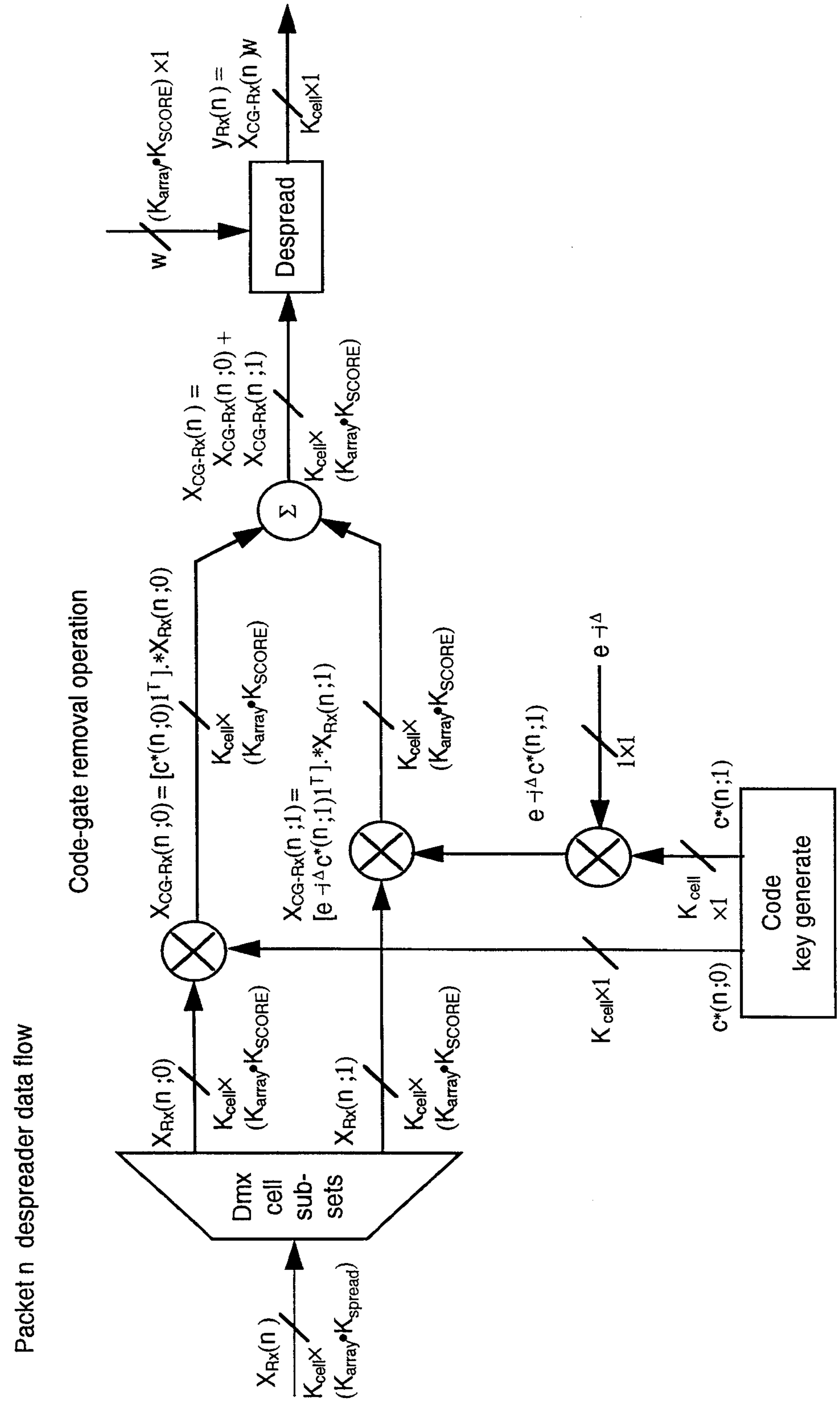
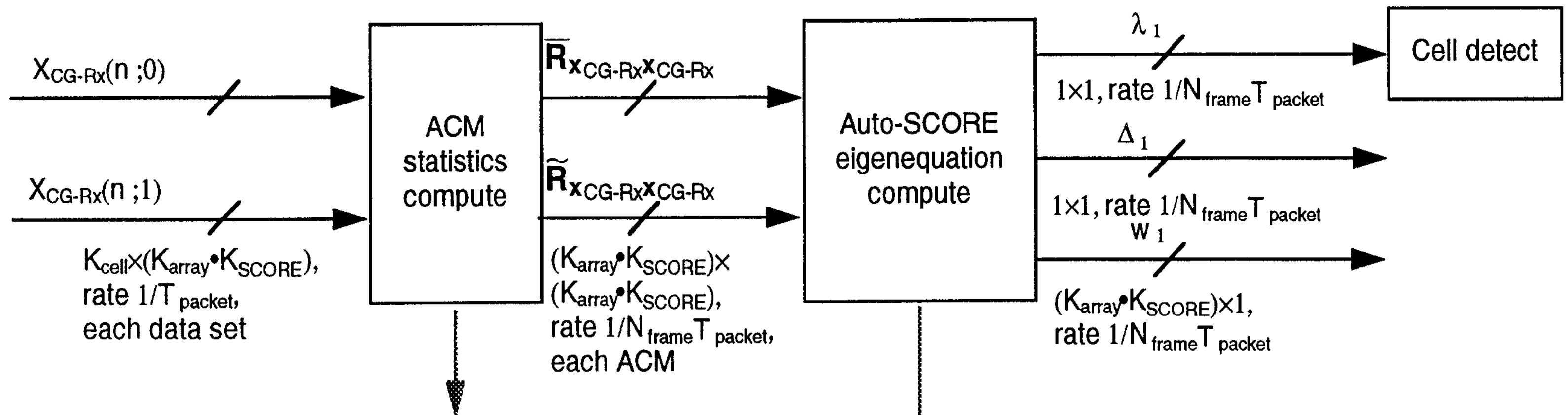

Packet n despreader data flow
Code-gate removal operation
$X_{Rx}(n)$
$K_{cell} \times (K_{array} \bullet K_{spread})$
Dmx cell sub-sets
$X_{Rx}(n;0)$
$K_{cell} \times (K_{array} \bullet K_{SCORE})$
$X_{Rx}(n;1)$
$K_{cell} \times (K_{array} \bullet K_{SCORE})$
$X_{CG\text{-}Rx}(n;0) = [c^*(n;0)1^T] .* X_{Rx}(n;0)$
$X_{CG\text{-}Rx}(n;1) = [e^{-j\Delta} c^*(n;1)1^T] .* X_{Rx}(n;1)$
$K_{cell} \times (K_{array} \bullet K_{SCORE})$
$K_{cell} \times 1$
$K_{cell} \times 1$
$e^{-j\Delta} c^*(n;1)$
$e^{-j\Delta}$
1×1
$c^*(n;0)$
$c^*(n;1)$
Code key generate
Σ
$X_{CG\text{-}Rx}(n) = X_{CG\text{-}Rx}(n;0) + X_{CG\text{-}Rx}(n;1)$
$K_{cell} \times (K_{array} \bullet K_{SCORE})$
w
$(K_{array} \bullet K_{SCORE}) \times 1$
Despread
$y_{Rx}(n) = X_{CG\text{-}Rx}(n)w$
$K_{cell} \times 1$

Fig. 33



$$\bar{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(k,l) = \frac{1}{K_{cell}N_{frame}} \sum_{q=0}^{N_{frame}-1} \mathbf{X}^{H}_{CG\text{-}Rx}(N_{frame}n+q;k)\,\mathbf{x}_{CG\text{-}Rx}(N_{frame}n+q;l)$$

$$\bar{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}} = \frac{1}{2}\left(\bar{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(0,0)+\bar{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(1,1)\right)$$

$$\tilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}} = \frac{1}{2}\left(\bar{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(0,1)+\bar{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(1,0)\right)$$

$$\lambda_q \bar{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}\mathbf{w}_q = \tilde{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}\mathbf{w}_q,\ \lambda_q \geq \lambda_{q+1}$$

$$\mathbf{w}_q^H \bar{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}\mathbf{w}_q = 1 \text{ (unit output power constraint)}$$

$$\Delta_q = \Delta_q + \arg\{\mathbf{w}_q^H \bar{\mathbf{R}}_{\mathbf{x}_{CG\text{-}Rx}\mathbf{x}_{CG\text{-}Rx}}(1,0)\mathbf{w}_q\}$$

Fig. 34

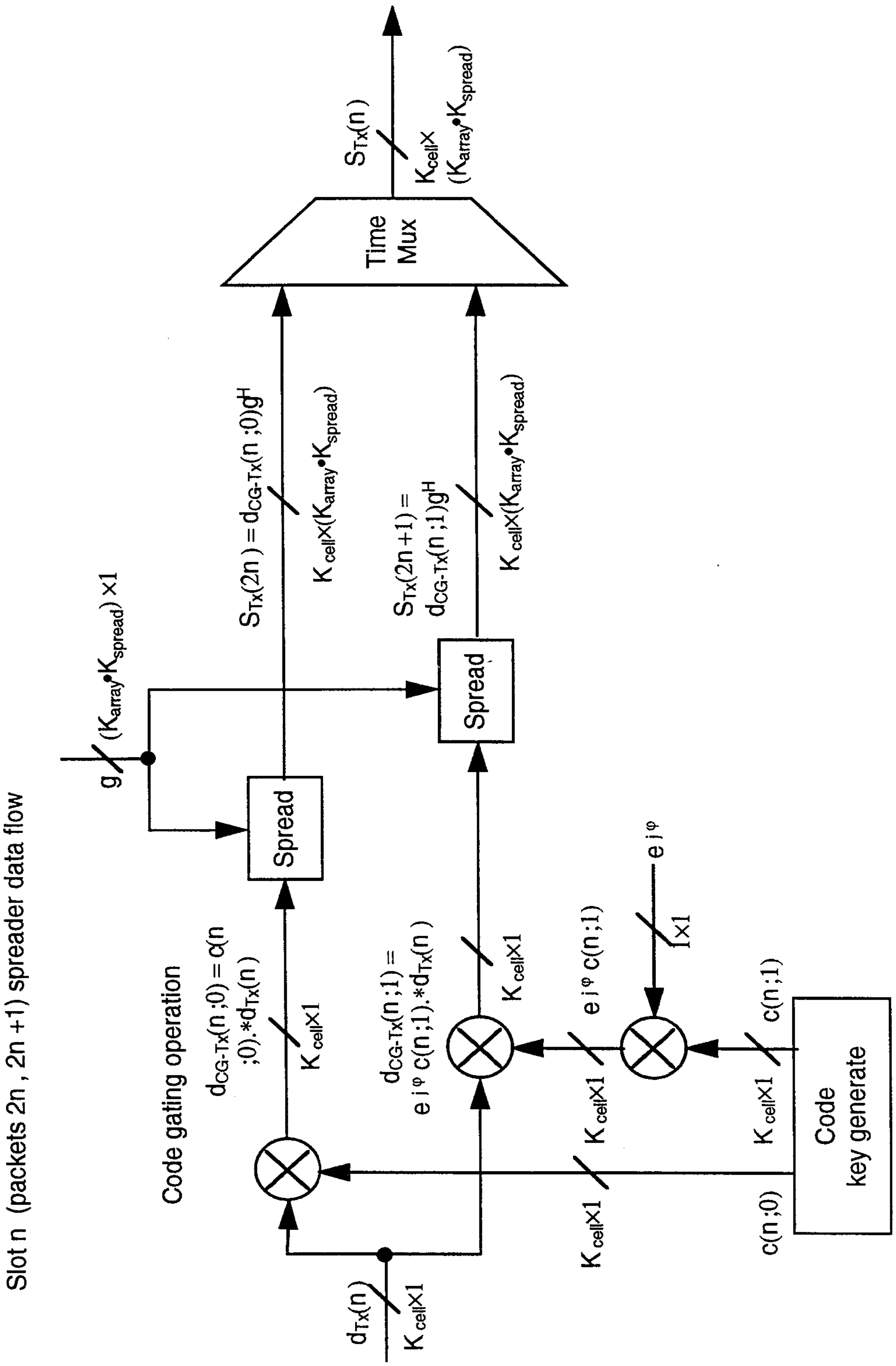
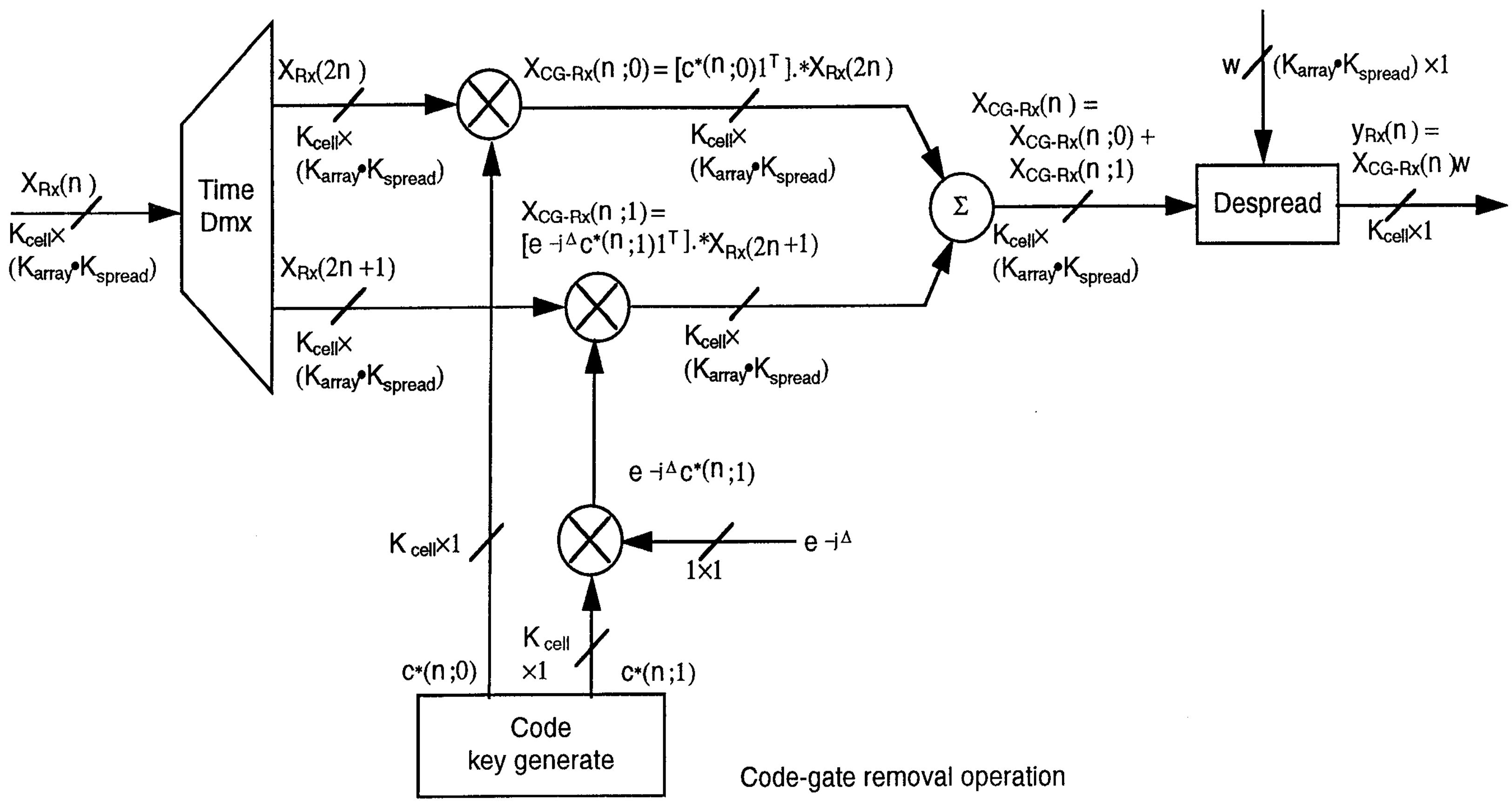

Slot n (packets 2n , 2n +1) spreader data flow
Code gating operation
$d_{CG-TX}(n;0) = c(n;0).*d_{TX}(n)$
$d_{CG-TX}(n;1) = e^{j\varphi} c(n;1).*d_{TX}(n)$
$d_{TX}(n)$
$K_{cell}\times 1$
g
$(K_{array}\bullet K_{spread})\times 1$
Spread
Spread
$S_{TX}(2n) = d_{CG-TX}(n;0)g^H$
$K_{cell}\times(K_{array}\bullet K_{spread})$
$S_{TX}(2n+1) = d_{CG-TX}(n;1)g^H$
$K_{cell}\times(K_{array}\bullet K_{spread})$
Time Mux
$S_{TX}(n)$
$K_{cell}\times(K_{array}\bullet K_{spread})$
$e^{j\varphi}$
1×1
$e^{j\varphi} c(n;1)$
$c(n;1)$
$c(n;0)$
Code key generate

Fig. 35

Slot n (packets 2n , 2n +1) despreader data flow



Code-gate removal operation

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US98/17063

A. CLASSIFICATION OF SUBJECT MATTER

IPC(6) : H04J 9/00; H04L 5/04; H04Q 7/00; H04B 1/06, 7/00, 7/216, 7/208;
US CL :Please See Extra Sheet.
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U.S. : 370/204, 310, 330, 335, 342, 344; 375/200, 206, 347; 455/270, 561

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched
NONE

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)
APS
search terms: cdma, tdm, tdma, fdma, diversity, spatial, space-division?, frequencies

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y	US 5,584,057 A (DENT) 10 DECEMBER 1996, column 5, line 59-column 6, line 63	1-86
Y	US 5,592,471 A (BRISKMAN) 07 JANUARY 1997, columns 2-8	1-86

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

* Special categories of cited documents:

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Date of the actual completion of the international search	Date of mailing of the international search report
04 NOVEMBER 1998	12 APR 1999

Name and mailing address of the ISA/US	Authorized officer
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A. CLASSIFICATION OF SUBJECT MATTER:
US CL :

370/204, 310, 330, 335, 342, 344; 375/200, 206, 347; 455/270, 561

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WORLD INTELLECTUAL PROPERTY ORGANIZATION
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INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(51) International Patent Classification 7 : H04J 13/00, H04Q 7/30, H04B 7/26

A1

(11) International Publication Number: WO 00/62456

(43) International Publication Date: 19 October 2000 (19.10.00)

(21) International Application Number: PCT/KR00/00345

(22) International Filing Date: 12 April 2000 (12.04.00)

(30) Priority Data:

1999/13610	12 April 1999 (12.04.99)	KR
1999/19080	26 May 1999 (26.05.99)	KR
1999/27355	7 July 1999 (07.07.99)	KR
1999/27398	8 July 1999 (08.07.99)	KR

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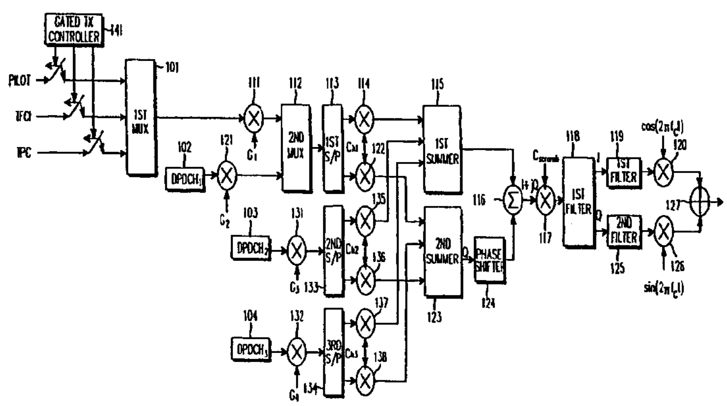
(81) Designated States: AU, BR, CA, CN, IL, IN, JP, RU, European patent (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE).

Published

With international search report.

(54) Title: APPARATUS AND METHOD FOR GATED TRANSMISSION IN A CDMA COMMUNICATION SYSTEM



(57) Abstract

A method and device for gating the transmission of dedicated control channels in a CDMA communication system is disclosed. In the device and method, if a base station (mobile station) determines whether there is no data to transmit to a mobile station (base station) for predetermined period of time, the base station (mobile station) gates transmission of control information according to a predetermined pattern on a dedicated control channel, which is used for transmitting control information to the mobile station (base station). Control information transmitted from the base station to the mobile station includes Transport Format Combination Indicator (TFCI), Transmit Power Control (TPC), and a pilot symbol. Control information transmitted from the mobile station to the base station includes TFCI, TPC, a pilot symbol, and FeedBack Information (FBI) for information about a transmit diversity antenna system. In a downlink DPCCH, transmission of the TFCI, TPC and pilot symbol of the predetermined n slots out of total slots of frame can be transmitted discontinuously during gated transmission. Alternatively, transmission of a pilot symbol of the predetermined nth slots and TFCI and TPC of (n+1)th slots can be transmitted discontinuously. In an uplink DPCCH, transmission of all the TFCI, TPC, FBI and pilot symbol of a specific slot are transmitted discontinuously during gated transmission. In addition, transmission of TPC for power control can be continuously transmitted for full rate power control when the traffic data is for transmission.

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APPARATUS AND METHOD FOR GATED TRANSMISSION IN A CDMA COMMUNICATION SYSTEM

BACKGROUND OF THE INVENTION

1. Field of the Invention

The present invention relates generally to a CDMA mobile communication system, and in particular, to an apparatus and method for gated transmission which does not require a separate resynchronization process by assigning dedicated channels.

2. Description of the Related Art

A conventional CDMA (Code Division Multiple Access) mobile communication system primarily provides a voice service. However, the future CDMA mobile communication system will support the IMT-2000 standard, which can provide high-speed data service as well as voice service. More specifically, the IMT-2000 standard can provide high-quality voice service, moving picture service, an Internet search service, etc.

In a mobile communication system, data communication is typically characterized by bursts of data transmissions alternating with long non-transmission periods. The bursts of data are referred to as "packets" or "packages" of data. In the future mobile communication system, traffic data is transmitted over a dedicated traffic channel for a data transmission duration, and the dedicated traffic channel is maintained for a predetermined time even when the base station and the mobile station have no traffic data to transmit. The mobile communication system, after finishing transmitting traffic data over the dedicated traffic channel,

maintains the down link and up link channels between the base station and the mobile station for a predetermined time even though there is no traffic data to transmit. This is done in order to minimize the time delay due to sync reacquisition when there is traffic data to transmit.

The invention will be described with reference to a UTRA (UMTS (Universal Mobile Telecommunications System) Terrestrial Radio Access) mobile communication system. Such a mobile communication system requires many states according to channel assignment circumstances and state information existence/nonexistence in order to provide a packet data service as well as a voice service. For example, a state transition diagram for a cell connected state, a user data active substate and a control-only substate are well defined in 3GPP RAN TS S2 series S2.03, 99. 04.

FIG. 1A shows state transition in the cell connected state of the mobile communication system. Referring to FIG. 1A, the cell connected state includes a paging channel (PCH) state, a random access channel (RACH)/downlink shared channel (DSCH) state, a RACH/forward link access channel (FACH) state, and a dedicated channel (DCH)/DCH, DCH/DCH+DSCH, DCH/DSCH+DSCH Ctrl (Control Channel) state.

FIG. 1B shows a user data active substate and a control-only substate of the DCH/DCH, DCH/DCH+DSCH, DCH/DSCH+DSCH Ctrl state. It should be noted that the novel gated transmission device and method is applied to a situation which has no taffic data for a predetermined time.

The existing CDMA mobile communication system, which mainly provides voice service, releases a channel after completion of data transmission and connects the channel again when there is further data to transmit. However, when providing packet data service as well as voice service, the conventional data transmission method has many delaying factors such as reconnection delay, thus making it difficult to provide high-quality service. Therefore, to provide packet data service as well as voice service, an improved data transmission method is required. For example, in many cases, data transmission is performed intermittently, such as for Internet access and file downloading. Therefore, there occurs a non-transmission period between transmissions of packet data. During this period, the

conventional data transmission method releases or maintains the dedicated traffic (or data) channel. If the dedicated traffic channel is released, a long time is required in order to reconnect the channel, and, if the dedicated traffic channel is maintained, channel resources are wasted and reverse power is wasted. To solve such problems, a dedicated control channel is provided between the base station and the mobile station so that for the data transmission period, a control signal related to the dedicated traffic channel is exchanged and for the non-transmission period, the dedicated traffic channel is released and only the dedicated control channel is maintained. Such a state is referred to as the "control-only substate".

A downlink (or forward link) for transmitting signals from the base station to the mobile station includes the following physical channels. A description of the physical channels which depart from the scope of the invention will be avoided for simplicity. The physical channels involved in the invention include a dedicated physical control channel (hereinafter, referred to as DPCCH) in which pilot symbols are included for sync acquisition and channel estimation, and a dedicated physical data channel (hereinafter, referred to as DPDCH) for exchanging traffic data with a specific mobile station. The downlink DPDCH includes the traffic data, and the downlink DPCCH includes, at each slot (or power control group), transport format combination indicator (hereinafter, referred to as TFCI) which is information about the format of transmission data, transmit power control (hereinafter, referred to as TPC) information which is a power control command, and control information such as the pilot symbols for providing a reference phase so that a receiver (the base station or the mobile station) can compensate the phase. The DPDCH and the DPCCH are time multiplexed within one power control group in down link, and the DPDCH and the DPCCH are separated by orthogonal codes each other in up link.

For reference, the invention will be described with reference to the case where the frame length is 10msec and each frame includes 16 power control groups, i.e., each power control group has a length of 0.625msec. Alternatively, the invention will also be described with reference to another case where the frame length is 10msec and each frame includes 15 power control groups, i.e., each power control group has a length of 0.667msec. It will be assumed herein that the power control group (0.625msec or 0.667msec) has the same time period as the slot (0.625msec or 0.667msec). The power control group (or slot) is comprised of pilot

symbol, traffic data, transmission data-related information TFCI, and power control information TPC in down link. The values stated above are given by way of example only.

FIG. 2A shows a slot structure including the downlink DPDCH and DPCCH. In FIG 2A, although the DPDCH is divided into traffic data 1 (Data1) and traffic data 2 (Data2), there is a case where the traffic data 1 does not exist and only the traffic data 2 exists according to the types of the traffic data. Table 1 below shows the symbols constituting the downlink DPDCH/DPCCH fields, wherein the number of TFCI, TPC and pilot bits in each slot can vary according to a data rate and a spreading factor.

Unlike the downlink DPDCH and DPCCH, uplink DPDCH and DPCCH for transmitting signals from the mobile station to the base station are separated by channel separation codes.

FIG. 2B shows a slot structure including the uplink DPDCH and DPCCH. In FIG. 2B, the number of TFCI, FBI, TPC and pilot bits can vary according to the service being provided (which changes the type of the traffic data), or because of transmit antenna diversity, or because of a handover circumstance. The FBI (FeedBack Information) is information that the mobile station requests about the antennas at the base station, when the base station uses the transmit diversity antennas. Tables 2 and 3 below show the symbols constituting the uplink DPDCH and DPCCH fields, respectively.

[Table 1] Downlink DPDCH/DPCCH Fields

Channel Bit Rate (kbps)	Channel Symbol Rate (ksps)	SF	Bits/Frame			Bits/Slot	DPDCH Bits/Slot		DPCCH Bits/Slot		
			DPDCH	DPCCH	TOT		N_{data1}	N_{data2}	N_{TFCI}	N_{TPC}	N_{pilot}
16	8	512	64	96	160	10	2	2	0	2	4
16	8	512	32	128	160	10	0	2	2	2	4
32	16	256	160	160	320	20	2	8	0	2	8
32	16	256	128	192	320	20	0	8	2	2	8
64	32	128	480	160	640	40	6	24	0	2	8
64	32	128	448	192	640	40	4	24	2	2	8
128	64	64	1120	160	1280	80	14	56	0	2	8
128	64	64	992	288	1280	80	6	56	8	2	8

256	128	32	2400	160	2560	160	30	120	0	2	8
256	128	32	2272	288	2560	160	22	120	8	2	8
512	256	16	4832	288	5120	320	62	240	0	2	16
512	256	16	4704	416	5120	320	54	240	8	2	16
1024	512	8	9952	288	10240	640	126	496	0	2	16
1024	512	8	9824	416	10240	640	118	496	8	2	16
2048	1024	4	20192	288	20480	1280	254	1008	0	2	16
2048	1024	4	20064	416	20480	1280	246	1008	8	2	16

[Table 2] Uplink DPDCH Fields

Channel Bit Rate (kbps)	Channel Symbol Rate (ksps)	SF	Bits/Frame	Bits/Slot	N_{data}
16	16	256	160	10	10
32	32	128	320	20	20
64	64	64	640	40	40
128	128	32	1280	80	80
256	256	16	2560	160	160
512	512	8	5120	320	320
1024	1024	4	10240	640	640

[Table 3] Uplink DPCCH Fields

Channel Bit Rate (kbps)	Channel Symbol Rate (ksps)	SF	Bits/Frame	Bits/Slot	N_{pilot}	N_{TPC}	N_{TFCI}	N_{FBI}
16	16	256	160	10	6	2	2	0
16	16	256	160	10	8	2	0	0
16	16	256	160	10	5	2	2	1
16	16	256	160	10	7	2	0	1
16	16	256	160	10	6	2	0	2
16	16	256	160	10	5	1	2	2

Tables 1 to 3 show an example where there exists one DPDCH which is a traffic channel, wherein SF denotes spreading factor. However, there may exist second, third and fourth DPDCHs according to the service types. Further, the downlink and uplink both may include several DPDCHs.

An exemplary hardware structure of the conventional mobile communication system (base station transmitter and mobile station transmitter) will be described below with reference to FIGS. 3A and 3B. Although the base station transmitter and mobile station transmitter will be described with reference to a case where there exist three DPDCHs, the number of DPDCHs is not limited.

FIG. 3A shows a structure of the conventional base station transmitter. Referring to FIG. 3A, multipliers 111, 121, 131 and 132 multiply a DPDCH signal and $DPDCH_1$, $DPDCH_2$ and $DPDCH_3$ signals, which have undergone channel encoding and interleaving, by gain coefficients G_1, G_2, G_3 and G_4, respectively. The gain coefficients G_1, G_2, G_3 and G_4 may have different values according to circumstances such as the service option and the handover. A multiplexer (MUX) 112 time-multiplexes the DPCCH signal and the $DPDCH_1$ signal into the slot structure of FIG. 2A. A first serial-to-parallel (S/P) converter 113 distributes the output of the multiplexer 112 to an I channel and a Q channel. Second and third S/P converters 133 and 134 S/P-convert the $DPDCH_2$ and $DPDCH_3$ signals and distribute them to the I channel and the Q channel, respectively. The S/P-converted I and Q channel signals are multiplied by channelization codes C_{ch1}, C_{ch2} and C_{ch3} in multipliers 114, 122, 135, 136, 137 and 138, for spreading and channel separation. Orthogonal codes are used for the channelization codes.

The I and Q channel signals multiplied by the channelization codes in the multipliers 114, 122, 135, 136, 137 and 138 are summed by first and second summers 115 and 123, respectively. That is, the I channel signals are summed by the first summer 115, and the Q channel signals are summed by the second summer 123. The output of the second summer 123 is phase shifted by 90° by a phase shifter 124. A summer 116 sums an output of the first summer 115 and an output of the phase shifter 124 to generate a complex signal I+jQ. A multiplier 117 scrambles the complex signal with a PN sequence C_{scramb} which is uniquely assigned to each base station, and a signal separator 118 separates the scrambled signal into a real part and an imaginary part and distributes them to the I channel and the Q channel. The I and Q channel outputs of the signal separator 118 are filtered by lowpass filters 119 and 125, respectively, to generate bandwidth-limited signals. The output signals of the filters 119 and 125 are multiplied by carriers $\cos\{2\pi f_c t\}$ and $\sin\{2\pi f_c t\}$ in multipliers 120 and 126, respectively, to frequency shift the signals to a radio frequency (RF) band. A summer 127 sums the frequency-shifted I and Q channel signals.

FIG. 3B shows a structure of the conventional mobile station transmitter. Referring to FIG. 3B, multipliers 211, 221, 223 and 225 multiply a DPCCH signal

and $DPDCH_1$, $DPDCH_2$ and $DPDCH_3$ signals, which have undergone channel encoding and interleaving, by channelization codes (orthogonal codes) C_{ch1}, C_{ch2}, C_{ch3} and C_{ch4}, respectively, for spreading and channel separation. Orthogonal codes are used for the channelization codes. The output signals of the multipliers 211, 221, 223 and 225 are multiplied by gain coefficients G_1, G_2, G_3 and G_4 in multipliers 212, 222, 224 and 226, respectively. The gain coefficients G_1, G_2, G_3 and G_4 may have different values. The outputs of the multipliers 212 and 222 are summed by a first summer 213 and output as an I channel signal, and the outputs of the multipliers 224 and 226 are summed by a second summer 227 and output as a Q channel signal. The Q channel signal output from the second summer 227 is phase shifted by 90° in a phase shifter 228.

A summer 214 sums the output of the first summer 213 and the output of the phase shifter 228 to generate a complex signal I+jQ. A multiplier 215 scrambles the complex signal with a PN sequence C_{scramb} which is uniquely assigned to each station, and a signal separator 229 separates the scrambled signal into a real part and an imaginary part and distributes them to the I channel and the Q channel. The I and Q channel outputs of the signal separator 229 are filtered by lowpass filters 216 and 230, respectively, to generate bandwidth-limited signals. The output signals of the filters 216 and 230 are multiplied by carriers $\cos\{2\pi f_c t\}$ and $\sin\{2\pi f_c t\}$ in multipliers 217 and 231, respectively, to frequency shift the signals to a radio frequency (RF) band. A summer 218 sums the frequency-shifted I and Q channel signals.

A conventional transmission signal structure of the base station and the mobile station will be made below. FIG. 5A shows how to transmit the downlink DPCCH and the uplink DPCCH when transmission of the uplink DPDCH is discontinued when there is no data to transmit for a predetermined time which is called control-only substate. FIG. 5B shows how to transmit the downlink DPCCH and the uplink DPCCH when transmission of the downlink DPDCH is discontinued when there is no data to transmit. As illustrated in FIGS. 5A and 5B, the mobile station constantly transmits the uplink DPCCH signal in spite of no DPDCH data in order to avoid a resynchronization acquisition process between the base station and the mobile station. When there is no traffic data to transmit for a long time, the base station and the mobile station make a transition to an RRC (Radio Resource

Control) connection release state (not shown in the FIGs.). In this state, transmission of the uplink DPCCH is discontinued, but the mobile station transmits pilot symbols and power control bits over the DPCCH until the transition is completed, thereby increasing interference in the uplink (or reverse link). The increase in interference of the uplink causes a decrease in the capacity of the uplink.

In the conventional method, although continuous transmission of the uplink DPCCH in the control-only substate is advantageous in that it is possible to avoid the sync reacquisition process in the base station, it creases an interference to the uplink and mobile station power consumption, causing a decrease in the capacity of the uplink. Further, in the downlink, continuous transmission of the uplink power control bits causes an increase in interference of the downlink and a decrease in the capacity of the downlink. Therefore, it is necessary to minimize the time required for the sync reacquisition process in the base station, to minimize the interference due to transmission of the uplink DPCCH and to minimize the interference and mobile station power consumption due to transmission of the uplink power control bits over the downlink.

SUMMARY OF THE INVENTION

It is, therefore, an object of the present invention to provide a communication device and method for minimizing the time required for a sync reacquisition process between base station and mobile station, for minimizing the interference and power consumption of mobile station due to transmission of a uplink DPCCH, and for minimizing the interference due to transmission of uplink power control bits over a downlink when there is no data to transmit on DPDCH for predetermined time.

It is another object of the present invention to provide a device and method for gating a dedicated control channel (DPCCH) signal on a gated transmission unit basis in a mobile communication system, wherein the gated transmission unit is either identical to an actual slot unit or different from the actual slot unit.

It is further another object of the present invention to provide a device and method for locating a power control bit in the last slot of each frame to control the

power of the first slot of the next frame in a mobile communication system.

To achieve the above object, a base station (or mobile station) according to the present invention determines whether there is data to transmit to the mobile station (or base station) on DPDCH. When there is no data to transmit on DPDCH, the base station (or mobile station) gates transmission of control information according to a predetermined time period pattern within one frame on a dedicated control channel. Here, "gated transmission" refers to transmitting the control information included in the DPCCH only at a specific power control group (PCG)/slot (or PCGs/slots) according to a predetermined time pattern. Control information transmitted from the base station to the mobile station includes TFCI information about a format of transmission data, TPC information for power control, and a pilot symbol. Control information transmitted from the mobile station to the base station includes TFCI information about a format of transmission data, TPC information for power control, a pilot symbol, and FBI information for requesting information about a phase difference between two antennas when the base station uses transmit diversity antenna. In a downlink DPCCH, the TFCI, TPC and pilot symbol in an n predetermined-power control group (or one slot) can be discontinuously transmitted in a frame during gated transmission. Alternatively, the pilot symbol in a predetermined nth power control group (or slot) and TFCI and TPC in (n+1)th power control group can be discontinuously transmitted in a frame. In an uplink DPCCH, the TFCI, TPC, FBI and pilot symbol in a specific power control group (or slot) are discontinuously transmitted during gated transmission. If there is a short data to transmit on DPDCH in gated transmission mode, the power control bit can be transmitted in all slot during transmit the short data. Further, a gating pattern for the downlink control information and a gating pattern for the uplink control information have an offset so that gating should occur at different time points.

BRIEF DESCRIPTION OF THE DRAWINGS

The above and other objects, features and advantages of the present invention will become more apparent from the following detailed description when taken in conjunction with the accompanying drawings in which:

FIG. 1A is a state transition diagram for a packet data service;

FIG. 1B is a state transition diagram between a user data active substate and a control-only substate of the DCH/DCH state;

FIG. 2A is a diagram illustrating a slot structure of downlink DPDCH and DPCCH;

FIG. 2B is a diagram illustrating a slot structure of uplink DPDCH and DPCCH;

FIG. 3A is a diagram illustrating a structure of a conventional base station transmitter;

FIG. 3B is a diagram illustrating a structure of a conventional mobile station transmitter;

FIG. 4A is a diagram illustrating a structure of a base station transmitter according to an embodiment of the present invention;

FIG. 4B is a diagram illustrating a structure of a mobile station transmitter according to an embodiment of the present invention;

FIG. 5A is a diagram for explaining how to transmit a downlink DPCCH and a uplink DPCCH when transmission of a uplink DPDCH is discontinued in a conventional control-only substate;

FIG. 5B is a diagram for explaining how to transmit a downlink DPCCH and a uplink DPCCH when transmission of a downlink DPDCH is discontinued in the conventional control-only substate;

FIG. 6A is a diagram illustrating a method for transmitting a signal according to a regular or gated transmission pattern for a uplink DPCCH according to an embodiment of the invention;

FIG. 6B is a diagram illustrating another method for transmitting a signal according to a regular or gated transmission pattern for a uplink DPCCH according to an embodiment of the invention;

FIG. 7A is a diagram illustrating a method for transmitting a signal when a uplink DPDCH message is generated while a uplink DPCCH is intermittently transmitted in a gating mode according to an embodiment of the invention;

FIG. 7B is a diagram illustrating another method for transmitting a signal when a uplink DPDCH message is generated while a uplink DPCCH is intermittently transmitted in a gating mode according to an embodiment of the invention;

FIG. 8A is a diagram illustrating a method for transmitting downlink and uplink signals when transmission of a downlink DPDCH is discontinued according to an embodiment of the present invention;

FIG. 8B is a diagram illustrating a method for transmitting downlink and uplink signals when transmission of a uplink DPDCH is discontinued according to an embodiment of the present invention;

FIG. 8C is a diagram illustrating another method for transmitting downlink and uplink signals when transmission of the downlink DPDCH is discontinued according to an embodiment of the present invention;

FIG. 8D is a diagram illustrating another method for transmitting downlink and uplink signals when transmission of the uplink DPDCH is discontinued according to an embodiment of the present invention;

FIG. 9A is a diagram illustrating a method for transmitting downlink and uplink signals when transmission of a downlink DPDCH is discontinued (gated transmission for the downlink DPCCH) according to an embodiment of the present invention;

FIG. 9B is a diagram illustrating a method for transmitting downlink and uplink signals when transmission of a uplink DPDCH is discontinued (gated transmission for downlink DPCCH) according to an embodiment of the present invention;

FIG. 10A is a diagram illustrating a structure of a base station transmitter according to another embodiment of the present invention;

FIG. 10B is a diagram illustrating a structure of a mobile station transmitter according to another embodiment of the present invention;

FIG. 11A is a diagram illustrating gated transmission for downlink and uplink DPCCHs according to a first embodiment of the present invention;

FIG. 11B is a diagram illustrating gated transmission for downlink and uplink DPCCHs according to a second embodiment of the present invention;

FIG. 11C is a diagram illustrating gated transmission for downlink and uplink DPCCHs according to a third embodiment of the present invention;

FIG. 11D is a diagram illustrating gated transmission for downlink and uplink DPCCHs according to a fourth embodiment of the present invention;

FIG. 11E is a diagram illustrating gated transmission for downlink and uplink DPCCHs according to a fifth embodiment of the present invention;

FIG. 12A is a diagram illustrating gated transmission for downlink and uplink DPCCHs according to a sixth embodiment of the present invention;

FIG. 12B is a diagram illustrating gated transmission for downlink and uplink DPCCHs according to a seventh embodiment of the present invention;

FIG. 12C is a diagram illustrating gated transmission for downlink and

uplink DPCCHs according to a eighth embodiment of the present invention;

FIG. 12D is a diagram illustrating gated transmission for downlink and uplink DPCCHs according to a ninth embodiment of the present invention; and

FIG. 12E is a diagram illustrating gated transmission for downlink and uplink DPCCHs according to a tenth embodiment of the present invention.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENT

Preferred embodiments of the present invention will be described herein below with reference to the accompanying drawings. In the following description, well-known functions or constructions are not described in detail since they would obscure the invention in unnecessary detail.

The term "normal transmission" as used herein refers to continuously transmitting control information included in the downlink or uplink DPCCH, i.e., TFCI, TPC and pilot symbols. Further, the term "gate transmission" refers to transmitting the control information included in the downlink DPCCH, i.e., TFCI, TPC and pilot symbols, only at a specific power control group (or slot) according to a predetermined time pattern. In addition, the term "gate transmission" refers to transmitting the control information included in the uplink DPCCH (i.e., TFCI, TPC and pilot symbols) only at a specific power control group (or one slot) according to a predetermined time pattern. The information, transmission of which is discontinued in the downlink DPCCH during gated transmission, may include all of the TFCI, TPC and pilot symbols in a predetermined nth power control group (or slot), or may include the pilot symbols in a predetermined nth power control group (or slot), and TFCI and TPC in an (n+1)th power control group. The information, transmission of which is discontinued in the uplink DPCCH during gated transmission, includes all of TFCI, TPC, FBI and pilot symbols in a specific power control group (or one slot). Herein, "a gated transmission unit is identical to a slot unit" means that TFCI, TPC and pilot symbols within one power control group are set as a gated transmission unit. Further, "a gated transmission unit is not identical to a slot unit" means that a pilot symbol in a predetermined nth slot and a TFCI and TPC in an (n+1)th slot are set as a gated transmission unit.

In addition, since performance at the beginning of a frame is very important, the invention locates the TPC for controlling the power of the first slot

of the next frame at the last slot of one frame. That is, TPC bits for the downlink DPCCH and the uplink DPCCH are located at the last slot of the nth frame, and power of the first slot of the (n+1)th frame is controlled using the TPC bits existing at the last slot of the nth frame.

Further, a power control rate can be maintained normal transmission even when transmission data is generated during gated transmission of the DPCCH signal according to the present invention. In addition, the gating pattern (or gated transmission pattern) for the downlink DPCCH and the gating pattern for the uplink DPCCH are determined to have an offset. That is, the control information for the downlink DPCCH and the control information for the uplink DPCCH are transmitted at different time points.

A hardware structure according to an embodiment of the invention will be described below.

FIG. 4A shows a structure of a base station transmitter according to an embodiment of the present invention. The base station transmitter is different from the conventional one of FIG. 3A in that with regard to the downlink DPCCH, the output of the multiplier 111 is gated by a gated transmission controller 141. That is, the gated transmission controller 141 performs gated transmission on the TFCI and TPC bits for the downlink DPCCH at a power control group (or time slot) scheduled with the mobile station when the traffic data is not transmitted over the downlink and uplink DPDCHs for predetermined time. In addition, the gated transmission controller 141 performs gated transmission on one power control group (or one entire slot) including the pilot symbols, TFCI and TPC bits for the downlink DPCCH at a power control group (or time slot) scheduled with the mobile station when the traffic data is not transmitted over the downlink and uplink DPDCHs for predetermined time.

Although the downlink gated transmission pattern is identical to the uplink gated transmission pattern, an offset may exist between them for efficient power control. The offset is given as a system parameter.

The gated transmission controller 141 can perform gated transmission either when the gated transmission unit is identical to the slot unit or when the

gated transmission unit is not identical to the slot unit. When the gated transmission unit is not identical to the slot unit, the gated transmission controller 141 separately gates the TFCI, TPC and pilot symbols. That is, the pilot symbol in the predetermined nth slot, and the TFCI and TPC in the (n+1)th slot are set as a gated on transmission unit.

In addition, the gated transmission controller 141 locates the TPC bits for power controlling the first slot of the next frame at the last slot of one frame to guarantee performance of the beginning part of the next frame. That is, the TPC bits for the downlink DPCCH and the uplink DPCCH are located at the last slot of the nth frame, and power of the first slot of the (n+1)th frame is controlled using the TPC bits existing at the last slot of the nth frame.

FIG. 4B shows a structure of a mobile station transmitter according to an embodiment of the present invention. The mobile station transmitter is different from the conventional one of FIG. 3B in that a gated transmission controller 241 is provided to gate transmission of the uplink DPCCH. That is, the gated transmission controller 241 performs gated transmission on one power control group (or one entire slot) including the pilot symbols, TFCI, FBI and TPC bits for the uplink DPCCH at a power control group (or time slot) scheduled with the mobile station in the control-only substate where the traffic data is not transmitted over the downlink and uplink DPDCHs. For sync detection, it is necessary to transmit the pilot symbols and TPC bits over the uplink DPCCH, and there is no alternative way to transmit the TPC, FBI and pilot symbols over the other uplink channels at the duration where transmission of the above channel is discontinued.

Now, a description will be made of a transmission signal structure of the base station and the mobile station according to an embodiment of the present invention.

FIG. 6A shows a method for transmitting a signal according to a regular or gated transmission pattern for the uplink DPCCH in case of there is no DPDCH data for a predetermined period of time according to an embodiment of the present invention. In FIG. 6A, reference numerals 301, 302, 303 and 304 show different gating rates according to a ratio of a duty cycle (hereinafter, referred to as DC). Reference numeral 301 shows a conventional method for transmitting the uplink

DPCCH without gating (DC=1, regular transmission), and reference numeral 302 shows a method for regularly transmitting every other power control group (or time slot), when DC=1/2 (only 1/2 of all the power control groups in one frame are transmitted). Reference numeral 303 shows a method for regularly transmitting every fourth power control group (3rd, 7th, 11th and 15th power control groups), when DC=1/4 (only 1/4 of all the power control groups in one frame are transmitted). Reference numeral 304 shows a method for regularly transmitting every eighth power control group (7th and 15th power control groups), when DC=1/8 (only 1/8 of all the power control groups in one frame are transmitted). In the embodiment of FIG. 6A, when DC=1/2 and 1/4, although the gated transmission controller 241 of the mobile station regularly gates the power control groups of the uplink DPCCH, it is also possible to gate arbitrary power control groups out of all the standard power control groups according to the corresponding DC. That is, when DC=1/2, it is also possible to gate arbitrary power control groups according to an irregular pattern, rather than to regularly transmit every other power control group. Further, when DC=1/2, it is also possible to continuously transmit half of all the power control groups at the second half (8th to 15th power control groups) of the frame. When DC=1/4, it is also possible to continuously transmit 1/4 of all the power control groups beginning at a 3/4 point of the frame (i.e., 12th to 15th power control groups). When DC=1/8, it is also possible to continuously transmit 1/8 of all the power control groups beginning at a 7/8 point of the frame (i.e., 14th to 15th power control groups).

The above gating rate transition method can be divided into several methods as stated below, and is determined according to system setup. In one method, a direct gating rate transition occurs from DC=1/1 to DC=1/2, from DC=1/1 to DC=1/4, or from DC=1/1 to DC=1/8 according to a set timer value or a transition command message from the base station. In another method, a sequential gating rate transition occurs from DC=1/1 to DC=1/2, from DC=1/2 to DC=1/4, or from DC=1/4 to 1/8. Selection of the DC value can be determined in consideration of the capacity of the corresponding mobile station or the quality of the channel environment.

FIG. 6B shows a method for transmitting a signal according to a regular or gated transmission pattern for the uplink DPCCH in case of there is no DPDCH data for a predetermined period of time according to another embodiment of the

present invention. In FIG. 6B, reference numerals 305, 306 and 307 show different gating rates according to a ratio of a duty cycle DC. Reference numeral 305 shows a method for transmitting two consecutive power control groups at regular locations (2^{nd}-3^{rd}, 6^{th}-7^{th}, 10^{th}-11^{th} and 14^{th}-15^{th} power control groups), when DC=1/2 (only 1/2 of all the power control groups in one frame are transmitted). Reference numeral 306 shows a method for transmitting two consecutive power control groups at regular locations (6^{th}-7^{th} and 14^{th}-15^{th} power control groups), when DC=1/4 (only 1/4 of all the power control groups in one frame are transmitted). Reference numeral 307 shows a method for transmitting two consecutive power control groups at regular locations (14^{th}-15^{th} power control groups), when DC=1/8 (only 1/8 of all the power control groups in one frame are transmitted). In the embodiment of FIG. 6B, when DC=1/2 and 1/4, although the gated transmission controller 241 of the mobile station regularly gates the power control groups of the uplink DPCCH, it is also possible to gate arbitrary power control groups out of all the power control groups according to the corresponding DC. That is, when DC=1/2, it is also possible to continuously gate 4 consecutive power control groups (e.g., 2^{nd}-5^{th} power control groups) according to an irregular pattern, rather than to regularly transmit every other 2 consecutive power control groups.

The above gating rate transition method can be divided into several methods as stated below, and is determined according to system setup. In one method, a direct state transition occurs from DC=1/1 (full rate) to DC=1/2, from DC=1/1 to DC=1/4, or from DC=1/1 to DC=1/8 according to a set timer value or a transition command message from the base station. In another method, a sequential gating rate transition occurs from DC=1/1 to DC=1/2, from DC=1/2 to DC=1/4, or from DC=1/4 to 1/8. Selection of the DC value can be determined in consideration of the capacity of the corresponding mobile station or the quality of the channel environment.

FIGS. 7A and 7B show the uplink DPCCH for the case where a transition message is transmitted over the uplink DPDCH when a dedicated MAC (Medium Access Control) logical channel is generated in case of there is nok DPDCH data for a predetermined period of time of FIGS. 6A and 6B. Reference numeral 311 of FIG. 7A shows a case where a uplink DPDCH message is generated while the uplink DPCCH does not undergo gated transmission (i.e., while the uplink DPCCH

is continuously transmitted (DC=1/1)). Reference numeral 312 shows a case where the uplink DPDCH message is generated while the uplink DPCCH undergoes DC=1/2 gated transmission. Reference numeral 313 shows a case where the uplink DPDCH message is generated while the uplink DPCCH undergoes DC=1/4 gated transmission. Reference numeral 314 shows a case where the uplink DPDCH message is generated while the uplink DPCCH undergoes DC=1/8 gated transmission.

The power control groups, as shown by the reference numerals 312, 313 and 314, are transmitted according to the gated transmission patterns in the first frame, and then undergo normal transmission when the uplink DPDCH is transmitted in the second frame. In the power control groups for normal transmission, the TPC bits for downlink power control can be omitted and the pilot duration (or period) can be extended to a power control group length. Beginning at the power control groups succeeding after transmitting the uplink DPDCH message by normal transmission of the power control groups, it is possible to transmit the uplink DPCCH without gating, or it is possible to gate transmission of the uplink DPCCH according to the original DC value until a gating rate transition message is received from the base station. That is, when the uplink DPDCH message is transmitted for DC=1/2 gated transmission, it is possible to perform normal transmission on the power control group of the above duration, thereafter perform DC=1/2 gated transmission again, and then perform DC=1 (regular transmission) gated transmission when the DPDCH user data exist.

Like the uplink DPCCH, even in the downlink, when a downlink DPDCH message is generated during gated transmission for the DPCCH, the power control groups, which are transmitted according to the gated transmission pattern, undergo normal transmission for transmit the downlink DPDCH. In the power control groups for normal transmission, the TPC bits for downlink power control can be omitted and the pilot duration can be extended to a power control group length. Beginning at the power control groups succeeding after transmitting the downlink DPDCH message by normal transmission of the power control groups, it is possible to transmit the downlink DPCCH without gating, or it is possible to gate transmission of the downlink DPCCH according to the original DC value until a state transition request message is received from the mobile station. That is, when the downlink DPDCH message is transmitted for DC=1/2 gated transmission, it is

possible to perform normal transmission on the power control group of the above duration, thereafter perform DC=1/2 gated transmission again, and then perform DC=1 gated transmission when transmitting the DPDCH user data.

Reference numeral 315 of FIG. 7B shows a case where a uplink DPDCH message is generated while the uplink DPCCH undergoes DC=1/2 gated transmission. Reference numeral 316 shows a case where the uplink DPDCH message is generated while the uplink DPCCH undergoes DC=1/4 gated transmission. Reference numeral 317 shows a case where the uplink DPDCH message is generated while the uplink DPCCH undergoes DC=1/8 gated transmission. The power control groups, as shown by the reference numerals 315, 316 and 317, are transmitted according to the gated transmission patterns, and then undergo normal transmission for transmit the downlink DPDCH message. In the power control groups for normal transmission, the TPC bits for downlink power control can be omitted and the pilot duration (or period) can be extended to a power control group length. Beginning at the power control groups succeeding after transmitting the uplink DPDCH message by normal transmission of the power control groups, it is possible to transmit the uplink DPCCH without gating, or it is possible to gate transmission of the uplink DPCCH according to the original DC value until a state transition message is received from the base station. That is, when the uplink DPDCH message is transmitted for DC=1/2 gated transmission, it is possible to perform normal transmission on the power control group of the above duration, thereafter perform DC=1/2 gated transmission again, and then perform DC=1 gated transmission when transmitting the DPDCH user data.

It is also possible to simultaneously gate transmission of both the uplink DPCCH and the downlink DPCCH according to the same gating pattern. Beginning at the power control groups succeeding after transmitting the downlink DPDCH message by normal transmission of the power control groups, generated while gating transmission of the downlink DPCCH, it is possible to transmit the downlink DPCCH without gating, or it is possible to gate transmission of the downlink DPCCH according to the original DC value until a state transition request message is received from the mobile station. That is, when the downlink DPDCH message is transmitted for DC=1/2 gated transmission, it is possible to perform normal transmission on the power control group of the above duration, thereafter perform DC=1/2 gated transmission again, and then perform DC=1 gated transmission

when transmitting the DPDCH user data.

FIG. 8A shows a method for transmitting downlink and uplink signals when transmission of a downlink DPDCH is discontinued. When transmission of the downlink DPDCH is discontinued as shown by reference numeral 801 in the user data active substate where there exists no uplink DPDCH, the base station and the mobile station start gating transmission if a set timer value expires or a downlink DPDCH message for state transition is generated. Although FIG. 8A shows an embodiment where the message for start gating transmission is generated by the base station, it is also possible for the mobile station to send a gating-request message to the base station when there is no downlink and uplink DPDCH. While transmitting the downlink DPCCH in FIG. 8A, it is also possible to transmit all the TFCI, TPC and pilot symbols without gating. Since the TPC bits include meaningless TPC values determined by measuring power strength of the pilot symbols of the gated power control groups within the uplink DPCCH, the mobile station ignores the meaningless TPC values transmitted from the base station in order to perform uplink power control in consideration of the gating pattern for the uplink DPCCH, and performs transmission at the same transmission power as the transmission power for the previous power control group. Alternatively, while transmitting the downlink DPCCH in FIG. 8A, it is also possible to gate only the TFCI and TPC bits in the downlink DPCCH without gating the pilot symbols in the downlink DPCCH. At this point, the gating pattern is identical to a gating pattern for the uplink DPCCH of the mobile station. The power control group, in which the TPC bits in the downlink DPCCH are gated, refers to the TPC bits generated by measuring the pilot symbols corresponding to the gated power control group in the DPCCH transmitted from the mobile station.

Reference numeral 802 shows a situation where a message for gated transmission is generated by the base station and is transmitted to the mobile station over the downlink DPDCH. In this case, the mobile station, which has been gating transmission of the uplink DPCCH, can stop gated transmission upon receipt of the message for stop the gated transmission and perform normal transmission (DC=1) when uplink DPDCH data should be transmitted. Alternatively, the mobile station, which has been gating transmission of the uplink DPCCH, can continue gated transmission even after receipt of the message for stop the gated transmission, stop gated transmission at the stop time included in the gated transmission stop

message, and then perform normal transmission (DC=1).

FIG. 8B shows a method for transmitting downlink and uplink signals when transmission of a uplink DPDCH is discontinued. When transmission of the uplink DPDCH is discontinued as shown by reference numeral 803 in the user data active substate where there exists no downlink DPDCH, the base station and the mobile station make the gated transmission at a time point appointed (or scheduled) between them when a set timer value expires or after exchanging a gated transmission message. Although FIG. 8B shows an embodiment where the message for gated transmission is generated in the downlink DPDCH, the gated transmission message can also be generated in the uplink DPDCH of the mobile station. While transmitting the downlink DPCCH in FIG. 8B, it is also possible to transmit all the TFCI, TPC and pilot symbols without gating. Since the TPC bits include meaningless TPC values determined by measuring power strength of the pilot symbols of the gated power control groups within the uplink DPCCH, the mobile station ignores the meaningless TPC values transmitted from the base station in order to perform uplink power control in consideration of the gating pattern for the uplink DPCCH, and performs transmission at the same transmission power as the transmission power for the previous power control group. Alternatively, while transmitting the downlink DPCCH in FIG. 8B, it is also possible to gate only the TFCI and TPC bits in the downlink DPCCH without gating the pilot symbols in the downlink DPCCH. At this point, the gating pattern is identical to a gating pattern for the uplink DPCCH of the mobile station. The power control group, in which the TPC bits in the downlink DPCCH are gated, refers to the TPC bits generated by measuring the pilot symbols corresponding to the gated power control group in the DPCCH transmitted from the mobile station.

Reference numeral 804 shows a situation where a gated transmission message generated by the base station is transmitted to the mobile station over the downlink DPDCH. In this case, the mobile station, which has been gating transmission of the uplink DPCCH, can stop gated transmission upon receipt of the message for stop gated transmission and perform normal transmission (DC=1). Alternatively, the mobile station, which has been gating transmission of the uplink DPCCH, can continue gated transmission even after receipt of the message for stop gated transmission, stop gated transmission at the stop time included in the gated transmission stop message, and then perform normal transmission (DC=1).

FIG. 8C shows a method for transmitting downlink and uplink signals when transmission of a downlink DPDCH is discontinued. When transmission of the downlink DPDCH is discontinued as shown by reference numeral 805 in the user data active substate where there exists no uplink DPDCH, the base station and the mobile station start gated transmission if a set timer value expires or a downlink DPDCH message for start gated transmission is transmitted. Although FIG. 8C shows an embodiment where the message for gated transmission message is generated by the base station, it is also possible for the mobile station to send a gated transmission request message to the base station when there is no downlink and uplink DPDCH. While transmitting the downlink DPCCH in FIG. 8C, it is also possible to transmit all the TFCI, TPC and pilot symbols without gating. Since the TPC bits include meaningless TPC values determined by measuring the power strength of the pilot symbols of the gated power control groups within the uplink DPCCH, the mobile station ignores the meaningless TPC values transmitted from the base station in order to perform uplink power control in consideration of the gating pattern for the uplink DPCCH, and performs transmission at the same transmission power as the transmission power for the previous power control group. Alternatively, while transmitting the downlink DPCCH in FIG. 8C, it is also possible to gate only the TFCI and TPC bits in the downlink DPCCH without gating the pilot symbols in the downlink DPCCH. At this point, the gating pattern is identical to a gating pattern for the uplink DPCCH of the mobile station. The power control group, in which the TPC bits in the downlink DPCCH are gated, refers to the TPC bits generated by measuring the pilot symbols corresponding to the gated power control group in the DPCCH transmitted from the mobile station.

Reference numeral 806 shows a situation where a gated transmission message is generated by the mobile station and is transmitted to the base station over the uplink DPDCH. In this case, the mobile station, which has been gating transmission of the uplink DPCCH, can stop gated transmission after transmission of the gated transmission message over the uplink DPDCH and then perform normal transmission (DC=1). Alternatively, the mobile station, which has been gating transmission of the uplink DPCCH, can continue gated transmission even after receiving gated transmission stop message, stop gated transmission at the stop time included in the gated transmission stop message, and then perform normal transmission (DC=1).

FIG. 8D shows a method for transmitting downlink and uplink signals when transmission of a uplink DPDCH is discontinued. When transmission of the uplink DPDCH is discontinued as shown by reference numeral 807 in the user data active substate where there exists no downlink DPDCH for a predetermined period of time, the base station and the mobile station can start gated transmission at a time point appointed (or scheduled) between them when a set timer value expires or after exchanging a gated transmission message. Although FIG. 8D shows an embodiment where the message for gated transmission is generated in the downlink DPDCH, the gated transmission message can also be generated in the uplink DPDCH of the mobile station. While transmitting the downlink DPCCH in FIG. 8D, it is also possible to transmit all the TFCI, TPC and pilot symbols without gating. Since the TPC bits include meaningless TPC values determined by measuring power strength of the pilot symbols of the gated power control groups within the uplink DPCCH, the mobile station ignores the meaningless TPC values transmitted from the base station in order to perform uplink power control in consideration of the gating pattern for the uplink DPCCH, and performs transmission at the same transmission power as the transmission power for the previous power control group. Alternatively, while transmitting the downlink DPCCH in FIG. 8D, it is also possible to gate only the TFCI and TPC bits in the downlink DPCCH without gating the pilot symbols in the downlink DPCCH. At this point, the gating pattern is identical to a gating pattern for the uplink DPCCH of the mobile station. The power control group, in which the TPC bits in the downlink DPCCH are gated, refers to the TPC bits generated by measuring the pilot symbols corresponding to the gated power control group in the DPCCH transmitted from the mobile station.

Reference numeral 808 shows a situation where a gated transmission message generated by the mobile station is transmitted to the base station over the uplink DPDCH. In this case, the mobile station, which has been gating transmission of the uplink DPCCH, can stop gated transmission after transmission of the gated transmission message over the uplink DPDCH and then perform normal transmission (DC=1). Alternatively, the mobile station, which has been gating transmission of the uplink DPCCH, can continue gated transmission even after transmission of the gated transmission stop message, stop gated transmission at the stop time included in the gated transmission stop message, and then perform

normal transmission (DC=1).

FIG. 9A shows a method for transmitting downlink and uplink signals when transmission of a downlink DPDCH is discontinued. When transmission of the downlink DPDCH is discontinued, the base station and the mobile station can start gated transmission at a time point appointed between them if a set timer value expires or after exchanging a gated transmission message. FIG. 9A shows a case where a gating pattern for the downlink DPCCH 501 is identical to a gating pattern for the uplink DPCCH 503. Although FIG. 9A shows an embodiment where the gated transmission message is generated through the downlink DPDCH, the gated transmission message can also be generated through the uplink DPDCH of the mobile station.

FIG. 9B shows a method for transmitting downlink and uplink signals when transmission of a uplink DPDCH is discontinued. When transmission of the uplink DPDCH is discontinued, the base station and the mobile station make a state transition at a time point appointed between them if a set timer value expires or after exchanging a state transition message. FIG. 9B shows a case where a gating pattern for the downlink DPCCH is identical to a gating pattern for the uplink DPCCH. Although FIG. 9B shows an embodiment where the state transition message is generated through the downlink DPDCH, the state transition message can also be generated through the uplink DPDCH of the mobile station.

In the foregoing drawings and descriptions, the downlink and uplink frames have the same frame starting point. However, in the UTRA system, the starting point of the uplink frame is artificially delayed by 250μsec as compared with the starting point of the downlink frame. This is to make power control time delay become one slot (=0.625ms) in consideration of the propagation delay of the transmission signal when the cell radius is below 30km. Therefore, with due consideration of the artificial time delay between the downlink and uplink frame start time, the methods for transmitting the DPCCH signal according to gated transmission are shown by FIGS. 11A to 11E. FIGS. 10A and 10B show structures of the base station controller and the mobile station controller, respectively, which enable such gated transmission.

FIG. 10A shows a structure of the base station controller according to another embodiment of the present invention. The base station transmitter is different from FIG. 4A in that the pilot, TFCI and TPC bits constituting the downlink DPCCH can be separately gated according to different gating patterns by the gated transmission controller 141. That is, the gated transmission controller 141 performs gated transmission on the pilot, TFCI and TPC bits for the downlink DPCCH at a power control group (or time slot) scheduled with the mobile station in the control-only substate where the traffic data is not transmitted over the downlink and uplink DPDCHs. By using the gated transmission controller 141, it is also possible to assemble a pilot of a nth slot and TFCI and TPC bits of a (n+1)th slot into a gated transmission unit. When the base station transmits signaling data using the gated transmission controller 141 during gated transmission in the control-only substate, it is possible to avoid performing gated transmission on the pilot and TFCI at the duration where the signaling data is transmitted.

Alternatively, the gated transmission controller 141 can perform gated transmission on one power control group (or one entire slot) including the pilot symbols, TFCI and TPC bits for the downlink DPCCH at a power control group (or time slot) scheduled with the mobile station in the control-only substate when the traffic data is not transmitted over the downlink and uplink DPDCHs.

Although the downlink gated transmission pattern is identical to the uplink gated transmission pattern, there can exist an offset therebetween for efficient power control. The offset is given as a system parameter.

FIG. 10B shows a structure of the mobile station transmitter according to another embodiment of the present invention. The mobile station transmitter is different from FIG. 4B in that the pilot, TFCI, FBI and TPC bits constituting the uplink DPCCH can be separately gated according to different patterns by the gated transmission controller 241. The gated transmission controller 241 gates transmission of the pilot, FTCI, FBI and TPC bits for the uplink DPCCH at a power control group (or time slot) scheduled with the mobile station in the control-only substate when the traffic data is not transmitted over the downlink and uplink DPDCHs. When the base station transmits signaling data using the gated transmission controller 241 during gated transmission in the control-only substate, it is possible to avoid performing gated transmission on the pilot and TFCI at the

duration when the signaling data is transmitted.

Alternatively, the gated transmission controller 241 can perform gated transmission on one power control group (or one entire slot) including the pilot symbols, TFCI, FBI and TPC bits for the uplink DPCCH at a power control group (or time slot) scheduled with the mobile station in the control-only substate when the traffic data is not transmitted over the downlink and uplink DPDCHs.

Although the downlink gated transmission pattern is identical to the uplink gated transmission pattern, there can exist an offset therebetween for efficient power control. The offset is given as a system parameter.

FIGS. 11A to 11E and FIGS. 12A to 12E show signal transmission diagrams for gated transmission performed by the base station and the mobile station transmitters of FIGS. 10A and 10B. FIGS. 11A to 11E show how to perform gated transmission when the frame length is 10msec and each frame includes 16 power control groups, i.e., each power control group has a length of 0.625msec. FIGS. 12A to 12E show how to perform gated transmission when the frame length is 10msec and each frame includes 15 power control groups, i.e., each power control group has a length of 0.667msec.

FIG. 11A shows gated transmission for the downlink and uplink DPCCHs according to a first embodiment of the present invention. As shown in FIG. 11A, a gated transmission unit for the downlink DPCCH may not be a slot unit. That is, with regard to two adjacent slots, a pilot symbol of an predetermined nth slot and TFCI and TPC bits of an (n+1)th slot are set as a gated transmission unit for the downlink DPCCH because of the pilot symbol is used for channel estimatin to detect the TFCI and TPC. For example, when the gating rate is 1/2, a pilot symbol of slot number 0 and TFCI and TPC bits of slot number 1 are set as a gated transmission unit for the downlink DPCCH. When the gating rate is 1/4, a pilot symbol of slot number 2 and TFCI and TPC bits of slot number 3 are set as a gated transmission unit for the downlink DPCCH. When the gating rate is 1/8, a pilot symbol of slot number 6 and TFCI and TPC bits of slot number 7 are set as a gated transmission unit for the downlink DPCCH. Here, the gated transmission unit for the downlink DPCCH is set to be different from the actual slot unit, since an nth pilot symbol may be required in the receiver to demodulate the (n+1)th TPC

according to a demodulation method for the TPC signal.

When a signaling message is generated during such gated transmission, the signaling message is transmitted over the downlink or uplink DPDCH. Therefore, performance of the frame starting point is very important. In the invention, as shown in FIG. 11A, the TPC for the downlink DPCCH and the TPC for the uplink DPCCH are located at slot number 15 (i.e., the 16^{th} slot, which is the last slot of the nth frame), so that the first slot of the (n+1)th frame is power controlled using the TPC bits existing in the last slot of an nth frame. That is, the TPC for power controlling the first slot of the next frame is located at the last slot of the present frame.

Meanwhile, in the UTRA system stated above, an offset between the downlink and uplink frame start points is fixed to 250μsec. However, in gated transmission of the downlink and uplink DPCCHs, the offset value can be changed to an arbitrary value while the base station and the mobile station exchange a parameter for DPCCH gated transmission in the call setup process. The offset value is set to a proper value in consideration of propagation delay of the base station and the mobile station in the call setup process. That is, when the cell radius is over 30Km, the offset value can be set to a value larger than the conventional offset value of 250μsec for DPCCH gated transmission, and this value can be determined through experiments.

FIG. 11B shows gated transmission for the downlink and uplink DPCCHs according to a second embodiment of the present invention. FIG. 11B shows a case where transmission of the downlink DPCCH goes ahead of transmission of the uplink DPCCH during gated transmission, for the gating rates of 1/2, 1/4 and 1/8. The difference (i.e., offset) is designated by "DL-UL timing" for the gating rates of 1/2, 1/4 and 1/8.

Referring to FIG. 11B, with regard to two adjacent slots, a pilot symbol of the predetermined-nth slot and TFCI and TPC of the (n+1)th slot are set as a gated transmission unit for the downlink DPCCH. For example, for the gating rate 1/2, a pilot symbol of slot number 0 and TFCI and TPC of slot number 1 are set as a gated transmission unit for the downlink DPCCH. For the gating rate 1/4, a pilot

symbol of slot number 2 and TFCI and TPC of slot number 3 are set as a gated transmission unit for the downlink DPCCH. For the gating rate 1/8, a pilot symbol of slot number 6 and TFCI and TPC of slot number 7 are set as a gated transmission unit for the downlink DPCCH.

In addition, it is noted that the TPC for power controlling the first slot of the next frame is located at the last slot of the present frame. That is, the TPC for the downlink DPCCH and the TPC for the uplink DPCCH are both located at slot number 15 (i.e., the 16th slot).

FIG. 11C shows gated transmission for the downlink and uplink DPCCHs according to a third embodiment of the present invention. FIG. 11C shows a case where transmission of the uplink DPCCH goes ahead of transmission of the downlink DPCCH during gated transmission, for the gating rates of 1/2, 1/4 and 1/8.

Referring to FIG. 11C, with regard to two adjacent slots, a pilot symbol of the predetermined nth slot and TFCI and TPC of the (n+1)th slot are set as a gated transmission unit for the downlink DPCCH. For example, for the gating rate 1/2, a pilot symbol of slot number 1 and TFCI and TPC of slot number 2 are set as a gated transmission unit for the downlink DPCCH. For the gating rate 1/4, a pilot symbol of slot number 2 and TFCI and TPC of slot number 3 are set as a gated transmission unit for the downlink DPCCH. For the gating rate 1/8, a pilot symbol of slot number 6 and TFCI and TPC of slot number 7 are set as a gated transmission unit for the downlink DPCCH.

In addition, it is noted that the TPC for power controlling the first slot of the next frame is located at the last slot of the present frame. That is, the TPC for the downlink DPCCH and the TPC for the uplink DPCCH are both located at a slot number 15 (i.e., the 16th slot).

FIG. 11D shows gated transmission for the downlink and uplink DPCCHs according to a fourth embodiment of the present invention. FIG. 11D shows a case where for the gating rates of 1/2, 1/4 and 1/8, transmission of the downlink DPCCH goes ahead of transmission of the uplink DPCCH during gated transmission, and the downlink and uplink gating patterns are set to the same period.

Referring to FIG. 11D, with regard to two adjacent slots, a pilot symbol of the predetermined nth slot and TFCI and TPC of the (n+1)th slot are set as a gated transmission unit for the downlink DPCCH. For example, for the gating rate 1/2, a pilot symbol of slot number 0 and TFCI and TPC of slot number 1 are set as a gated transmission unit for the downlink DPCCH. For the gating rate 1/4, a pilot symbol of slot number 0 and TFCI and TPC of slot number 1 are set as a gated transmission unit for the downlink DPCCH. For the gating rate 1/8, a pilot symbol of slot number 2 and TFCI and TPC of slot number 3 are set as a gated transmission unit for the downlink DPCCH.

In addition, it is noted that the TPC for power controlling the first slot of the next frame is located at the last slot of the present frame. That is, the TPC for the downlink DPCCH and the TPC for the uplink DPCCH are both located at slot number 15 (i.e., the 16^{th} slot).

FIG. 11E shows gated transmission for the downlink and uplink DPCCHs according to a fifth embodiment of the present invention. FIG. 11E shows a case where for the gating rates of 1/2, 1/4 and 1/8, transmission of the uplink DPCCH goes ahead of transmission of the downlink DPCCH during gated transmission, and the downlink and uplink gating patterns are set to the same period.

Referring to FIG. 11E, with regard to two adjacent slots, a pilot symbol of the nth slot and TFCI and TPC of the (n+1)th slot are set as a gated transmission unit for the downlink DPCCH. For example, for the gating rate 1/2, a pilot symbol of slot number 1 and TFCI and TPC of slot number 2 are set as a gated transmission unit for the downlink DPCCH. For the gating rate 1/4, a pilot symbol of slot number 2 and TFCI and TPC of slot number 3 are set as a gated transmission unit for the downlink DPCCH. For the gating rate 1/8, a pilot symbol of slot number 6 and TFCI and TPC of slot number 7 are set as a gated transmission unit for the downlink DPCCH.

In addition, it is noted that the TPC for power controlling the first slot of the next frame is located at the last slot of the present frame. That is, the TPC for the downlink DPCCH and the TPC for the uplink DPCCH are both located at a slot number 15 (i.e., the 16^{th} slot).

FIG. 12A shows gated transmission for the downlink and uplink DPCCHs according to a sixth embodiment of the present invention. FIG. 12A shows a case where a gating rate for gated transmission of the downlink and uplink DPCCHs is 1/3, i.e., gated transmission is performed at the periods corresponding to 1/3 power control groups of the whole power control groups. That is, gated transmission is performed at the periods corresponding to 5 power control groups out of the whole 15 power control groups. At this point, a gated transmission unit for the downlink DPCCH is set to be different from a slot unit. That is, with regard to two adjacent slots, a pilot symbol of the predetermined nth slot and TFCI and TPC of the (n+1)th slot are set as a gated transmission unit for the downlink DPCCH because of the pilot symbol is used for channel estimation to detect the TFCI and the TPC.

In FIG. 12A, <Case 1> shows a case where the uplink DPCCH and the downlink DPCCH are transmitted at the same time during gated transmission, and the downlink and uplink gating patterns are set to the same period. With regard to two adjacent slots, a pilot symbol of slot number 1 and TFCI and TPC of slot number 2 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 4 and TFCI and TPC of slot number 5 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 7 and TFCI and TPC of slot number 8 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 10 and TFCI and TPC of slot number 11 are set as a gated transmission unit for the downlink DPCCH; and a pilot symbol of slot number 13 and TFCI and TPC of slot number 14 are set as a gated transmission unit for the downlink DPCCH.

<Case 2> shows a case where transmission of the uplink DPCCH occurs before transmission of the downlink DPCCH during gated transmission. At this point, with regard to two adjacent slots, a pilot symbol of slot number 0 and TFCI and TPC of slot number 1 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 3 and TFCI and TPC of slot number 4 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 6 and TFCI and TPC of slot number 7 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 9 and TFCI and TPC of slot number 10 are set as a gated transmission unit for the downlink DPCCH; and a pilot symbol of slot number 12 and TFCI and TPC of slot number 13 are set as a

gated transmission unit for the downlink DPCCH.

<Case 3> shows a case where transmission of the uplink DPCCH occurs before transmission of the downlink DPCCH during gated transmission. At this point, with regard to two adjacent slots, a pilot symbol of slot number 1 and TFCI and TPC of slot number 2 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 4 and TFCI and TPC of slot number 5 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 7 and TFCI and TPC of slot number 8 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 10 and TFCI and TPC of slot number 11 are set as a gated transmission unit for the downlink DPCCH; and a pilot symbol of slot number 13 and TFCI and TPC of slot number 14 are set as a gated transmission unit for the downlink DPCCH.

<Case 4> shows a case where transmission of the uplink DPCCH occurs after transmission of the downlink DPCCH during gated transmission. At this point, with regard to two adjacent slots, a pilot symbol of slot number 14 and TFCI and TPC of slot number 0 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 2 and TFCI and TPC of slot number 3 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 5 and TFCI and TPC of slot number 6 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 8 and TFCI and TPC of slot number 9 are set as a gated transmission unit for the downlink DPCCH; and a pilot symbol of slot number 11 and TFCI and TPC of slot number 12 are set as a gated transmission unit for the downlink DPCCH.

<Case 5> shows a case where transmission of the uplink DPCCH occurs after transmission of the downlink DPCCH during gated transmission. At this point, with regard to two adjacent slots, a pilot symbol of slot number 0 and TFCI and TPC of slot number 1 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 3 and TFCI and TPC of slot number 4 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 6 and TFCI and TPC of slot number 7 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 9 and TFCI and TPC of slot number 10 are set as a gated transmission unit for the downlink DPCCH; and a pilot symbol of slot number 12 and TFCI and TPC of a slot number 13 are set as a

gated transmission unit for the downlink DPCCH.

FIG. 12B shows gated transmission for the downlink and uplink DPCCHs according to a seventh embodiment of the present invention. FIG. 12A shows a case where the gating rate for gated transmission of the downlink and uplink DPCCHs is 1/5, i.e., gated transmission is performed so that 1/5 of the power control groups are transmitted in comparison to all the power control groups in standard transmission. That is, gated transmission is performed so that 3 power control groups out of the standard 15 power control groups are transmitted. At this point, a gated transmission unit for the downlink DPCCH is set to be different from a slot unit. That is, with regard to two adjacent slots, a pilot symbol of the predetermined nth slot and TFCI and TPC of the (n+1)th slot are set as a gated transmission unit for the downlink DPCCH because of the pilot symbol is used for channel estimation to detect the TFCI and the TPC.

Referring to FIG. 12B, with regard to two adjacent slots, a pilot symbol of slot number 3 and TFCI and TPC of slot number 4 are set as a gated transmission unit for the downlink DPCCH; a pilot symbol of slot number 8 and TFCI and TPC of slot number 9 are set as a gated transmission unit for the downlink DPCCH; and a pilot symbol of slot number 13 and TFCI and TPC of slot number 14 are set as a gated transmission unit for the downlink DPCCH.

FIG. 12C shows gated transmission for the downlink and uplink DPCCHs according to an eighth embodiment of the present invention. Referring to FIG. 12C, the gating pattern is set such that the last power control group of the uplink DPCCH should not be gated in the control-only substate. Such a gating pattern has high channel estimation performance, since the base station can perform channel estimation using the pilot symbols in the last power control group of the frame. In addition, it is possible to increase the time required when the base station processes the FBI bits transmitted from the mobile station.

FIG. 12D shows gated transmission for the downlink and uplink DPCCHs according to a ninth embodiment of the present invention. Shown is a gating pattern for transmitting a downlink message during gated transmission in the control-only substate.

Referring to FIG. 12D, for the frame period where the downlink message is transmitted (i.e., DPDCH transmission period), gated transmission is discontinued for the pilot and TFCI, and only the TPC continues to undergo gated transmission according to the gating pattern. For the period where the downlink data (message) is not transmitted, the pilot symbols and TFCI as well as TPC undergo gated transmission. The pilot symbol is transmitted at the 0th, 3rd, 6th, 9th and 12th slots only, and the TFCI and TPC bits are transmitted at the 1st, 4th, 7th, 10th and 13th slots only. When the downlink data is transmitted during such gated transmission, the pilot symbol and TFCI are transmitted at every slot, whereas TPC is transmitted at the 1st, 4th, 7th 10th and 13th slots only. Accordingly, even though downlink transmission data is generated during gated transmission, the power control rate is maintained.

FIG. 12E shows gated transmission for the downlink and uplink DPCCHs according to a tenth embodiment of the present invention. Shown is a gating pattern for transmitting a uplink message during gated transmission in the control-only substate. For a period where the uplink data (message) is not transmitted, the pilot symbols and TFCI as well as TPC and FBI undergo gated transmission. The pilot symbol, TFCI, FBI and TPC are transmitted at 2nd, 5th, 8th, 11th and 14th slots only. When the uplink data is transmitted during such gated transmission, the pilot symbol and TFCI are transmitted at every slot, whereas TPC and FBI are transmitted at the 2nd, 5th, 8th, 11th and 14th slots only. Accordingly, even though uplink transmission data is generated during gated transmission, the power control rate is maintained.

As shown in FIGS. 12D and 12E, for the DPDCH transmission period where the uplink message is transmitted, several embodiments of the invention discontinue gated transmission of the pilot and TFCI and continue to transmit FBI and TPC according to the gating rate.

As described above, the invention minimizes the time required for the sync reacquisition process in the base station, minimizes interference as well as uplink DPCCH transmission time, and minimizes interference due to the transmission of the uplink power control bit over the downlink, all of which increases the capacity of the mobile communication system.

While the invention has been shown and described with reference to certain preferred embodiments thereof, it will be understood by those skilled in the art that various changes in form and details may be made therein without departing from the spirit and scope of the invention as defined by the appended claims.

CLAIMS:

1. A method for transmitting down link DPCCH signals in a base station for a mobile communication system, comprising the steps of:

determining whether the base station has down link DPDCH data to transmit to a mobile station; and

gating transmission of the down link DPCCH signal according to a predetermined pattern when there is no data to transmit for predetermined period of time.

2. The method as claimed in claim 1, wherein the the DPCCH signal is transmitted in slot format, said slot having power control bit which control the up link transmission power , and the predetermined pattern is a pattern for gating transmission of the DPCCH slot signal during gated transmission of DPCCH signal.

3. The method as claimed in claim 1, wherein the DPCCH signal includes power control bit.

4. The method as claimed in claim 2, wherein the DPCCH signal includes a pilot symbol, a format of transmission data frame, and power control bit for up link transmission power control.

5. The method as claimed in claim 4, wherein the slot format includes pilot symbols, TFCI bits and power control bit, and predetermined pattern is a pattern for gating transmission of the pilot symbol, the TFCI bitsand the power control bit at predetermined n slots out of total slots of frame.

6. The method as claimed in claim 4, wherein the slot format includes pilot symbols, TFCI bits and power control bit, and predetermined pattern is a pattern for gating transmission of the pilot symbol at an predetermined nth slot and the TFCI bits and power control bitat a (n+1)th slot.

7. The method as claimed in claim 2, wherein the power control bit gating transmission is maintained regular when the base station transmit DPDCH data to the mobile station.

8. A base station transmission device for a mobile communication system, comprising:

a multiplexer for multiplexing a first channel signal and a second channel signal into a frame which is segmented into a plurality of slots, and for outputting the frame ;

a switch for gating the first channel signal; and

a controller for gating the switch such that the first channel signal undergoes gated transmission within a frame according to a predetermined pattern when there is no second channel signal to be transmitted to a mobile station.

9. The base station transmission device as claimed in claim 8, wherein the predetermined pattern is a pattern for gating transmission of the first channel signal at predetermined slots.

10. The base station transmission device as claimed in claim 8, wherein the first channel signal includes power control bit.

11. The base station transmission device as claimed in claim 9, wherein the first channel signal includes a pilot symbol, TFCI bits, and power control bit.

12. The base station transmission device as claimed in claim 11, wherein the predetermined pattern is a pattern for gating transmission of the pilot symbol, the TFCI bits and the power control bit at an predetermined n slot out of total slots of frame.

13. The base station transmission device as claimed in claim 11, wherein the predetermined pattern is a pattern for gating transmission of the pilot symbol at an predetermined nth slot and the TFCI bits and the power control bit at a (n+1)th slot.

14. The base station transmission device as claimed in claim 10, wherein the controller maintains regular transmission for the power control bit when the base station transmit DPDCH data to the mobile station during gated transmission.

15. The base station transmission device as claimed in claim 8, wherein the first channel is a DPCCH.

16. The base station transmission device as claimed in claim 8, wherein the second channel is a DPDCH.

17. A method for transmitting control DPCCH signal in a mobile station of a mobile communication system, comprising the steps of:

determining whether the mobile station has data to transmit to a base station; and

gating transmission of the DPCCH signal according to a predetermined pattern for maintain link power control loop when there is no data to transmit for predetermined period of time.

18 The method as claimed in claim 17, wherein the DPCCH signal have frame format, said frame includes a plurality of slots, and the predetermined pattern is a pattern for gating transmission of the DPCCH signal.

19. The method as claimed in claim 17, wherein the DPCCH signal includes at least power control bit.

20. The method as claimed in claim 18, wherein the DPCCH signal includes a pilot symbol, TFCI bits, and FBI bits for at least one phase difference between at least two antennas when the base station uses transmit diversity antennas.

21. The method as claimed in claim 20, wherein the predetermined pattern is a pattern for gating transmission of the pilot symbol, the TFCI bits, the power control bit, and FBI bits at predetermined slots.

22. The method as claimed in claim 19, wherein the up link DPCCH signal is transmitted continuously during mobile station transmit DPDCH data the base station.

23. The method as claimed in claim 22, wherein the transmission

power of the DPDCH data is increased than regular transmission.

24. A mobile station transmission device for a mobile communication system, comprising:

a Dedicated Physical Control Channel(DPCCH) for transmitting pilot ssymbol, TFCI bits for indicating frame format of Dedicated Physical Data Channel(DPDCH) frame, FBI bits for feed back information of diversity antenna signal, and power control bit for control down link transmission power

a Dedicated Physical Data Channel(DPDCH) for transmitting user data or signaling data to base station with the Dedicated Physical Control Channel(DPCCH)

a switch for gating the Dedicated Physical Control Channel(DPCCH) signal; and

a controller for gating the switch such that the Dedicated Physical Control Channel(DPCCH) signal undergoes gated transmission within the frame according to a predetermined pattern when there is no Dedicated Physical Data Channel(DPDCH) signal to be transmitted to the base station for predetermined period of time.

25. The mobile station transmission device as claimed in claim 24, wherein the predetermined pattern is a pattern for gating transmission of the Dedicated Physical Control Channel(DPCCH) signal at predetermined slots.

26. The mobile station transmission device as claimed in claim 24, wherein the Dedicated Physical Control Channel(DPCCH) signal includes power control information.

27. The mobile station transmission device as claimed in claim 25, wherein the Dedicated Physical Control Channel(DPCCH) and the Dedicated Physical Data Channel(DPDCH) is spread with orthogonal code respectively for channel separation and the channel signals are multiplied with gain value respectively.

28. The mobile station transmission device as claimed in claim 27, wherein the time period pattern is a pattern for gating transmission of the pilot symbol, theTFCI bits, the FBI bits and the power control bit at an predetermined

nth slot out of total slots of a frame.

29. The mobile station transmission device as claimed in claim 26, wherein the controller control the DPCCH signal with regularly when the mobile station transmit DPDCH data.

30. A method for transmitting control information in a mobile communication system, comprising the steps of:

(a) determining whether a base station has data to transmit to a mobile station;

(b) transmitting gating message to indicate gating start time and gating pattern to a mobile station when the data to transmit has not exist predetermined period of time;

(c) gating transmission of first control information according to gating pattern on a downlink dedicated control channel, said downlink dedicated control channel for transmitting the first control information to the mobile station;

(d) determining whether the mobile station has data to transmit to the base station;transmitting gating request message to the base station when the data to transmit has not exist predetermined period of time;

(e) gating, when the mobile station has received the gating message and reach the gating start time from the base station, transmission of second control information according to a predetermined second pattern within one frame on an uplink dedicated control channel, said uplink dedicated control channel for transmitting the second control information to the base station.

31. The method as claimed in claim 30, wherein the frame on the downlink dedicated control channel is segmented into a plurality of slots and the predetermined first pattern is a pattern for gating transmission of the first control information at predetermined slots.

32. The method as claimed in claim 30, wherein the first control information includes power control information.

33. The method as claimed in claim 31, wherein the first control information includes a pilot symbol, first information about a format of transmission data, and second information for power control.

34. The method as claimed in claim 33, wherein the predetermined first pattern is a pattern for gating transmission of the pilot symbol, the first information and the second information at an predetermined nth slot.

35. The method as claimed in claim 33, wherein the predetermined first pattern is a pattern for gating transmission of the pilot symbol at an predetermined nth slot and the first information and the second information at a (n+1)th slot.

36. The method as claimed in claim 31, wherein gated transmission for the power control information is maintained when the base station generates data to be transmitted to the mobile station during gated transmission of the first control information.

37. The method as claimed in claim 30, wherein the frame on the uplink dedicated control channel is segmented into a plurality of slots and the predetermined second pattern is a pattern for gating transmission of the second control information at predetermined slots.

38. The method as claimed in claim 37, wherein the second control information includes power control information.

39. The method as claimed in claim 37, wherein the second control information includes a pilot symbol, first information about a format of the transmission data, and second information for requesting information about at least one phase difference between at least two antennas when the base station uses transmit diversity antennas, and third information for power control.

40. The method as claimed in claim 37, wherein the predetermined second pattern is a pattern for gating transmission of the pilot symbol, the first information, the second information and the third information at predetermined slots.

41. The method as claimed in claim 38, wherein gated transmission for the power control information is maintained when the mobile station has data to

transmit to the base station during gated transmission of the second control information.

42. The method as claimed in claim 39, wherein gated transmission for the second information and the third information is maintained when the mobile station has data to transmit to the base station during gated transmission of the second control information.

43 The method as claimed in claim 30, wherein there is an time offset between the predetermined first pattern and the predetermined second pattern.

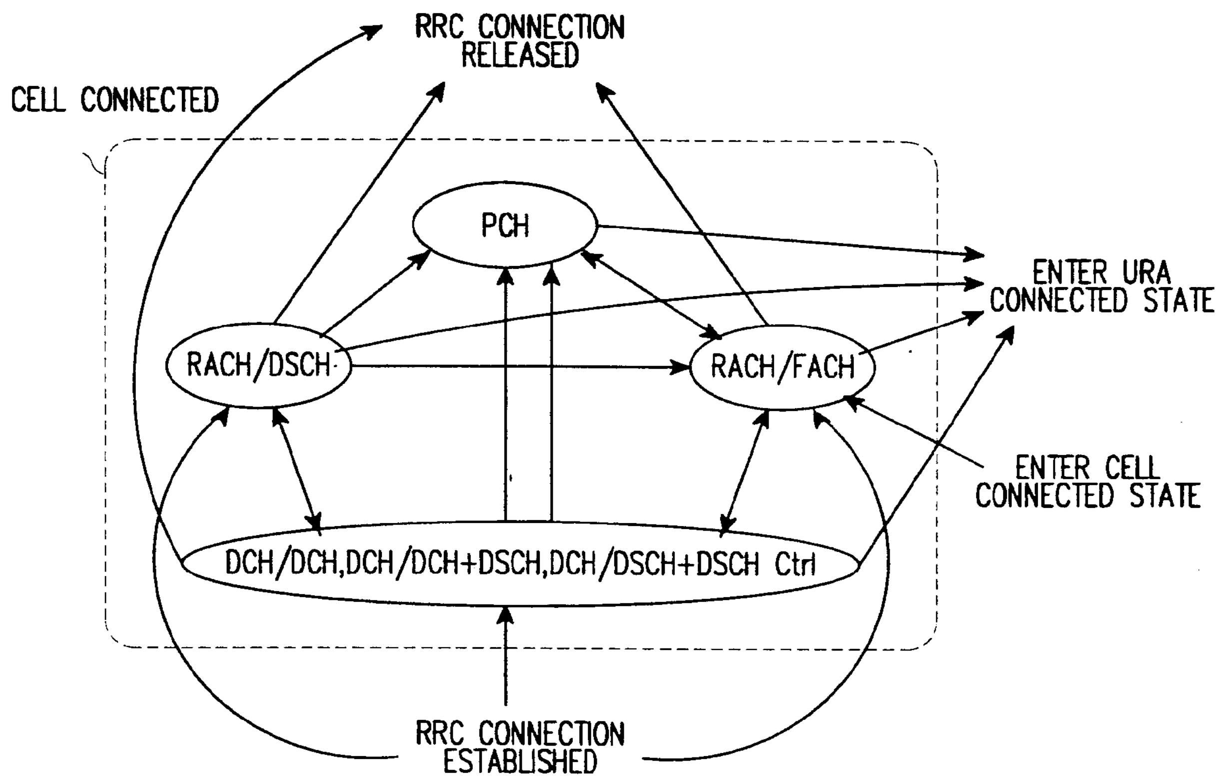

RRC CONNECTION RELEASED
CELL CONNECTED
PCH
ENTER URA CONNECTED STATE
RACH/DSCH
RACH/FACH
ENTER CELL CONNECTED STATE
DCH/DCH,DCH/DCH+DSCH,DCH/DSCH+DSCH Ctrl
RRC CONNECTION ESTABLISHED

FIG. 1A

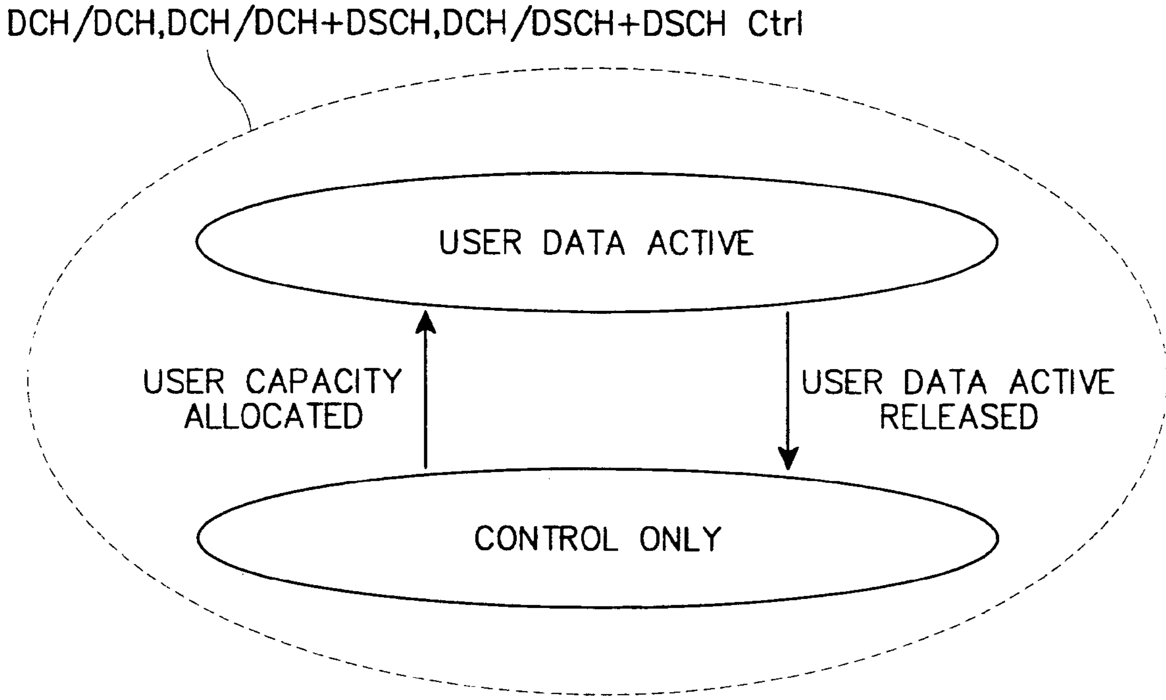

DCH/DCH,DCH/DCH+DSCH,DCH/DSCH+DSCH Ctrl
USER DATA ACTIVE
USER CAPACITY ALLOCATED
USER DATA ACTIVE RELEASED
CONTROL ONLY

FIG. 1B

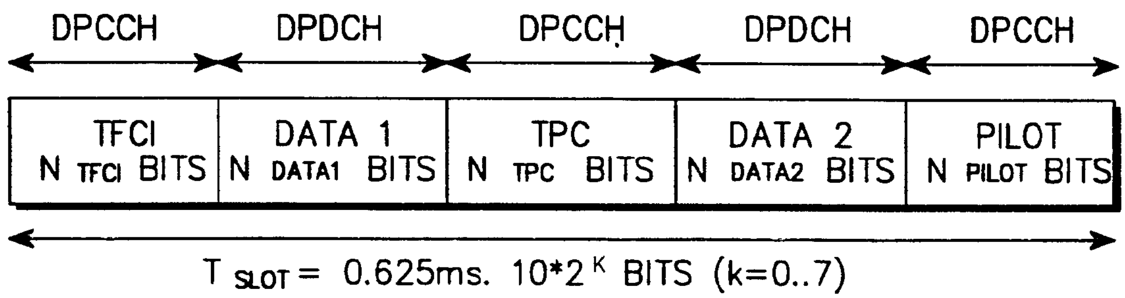

DPCCH
DPDCH
DPCCH
DPDCH
DPCCH
TFCI
N TFCI BITS
DATA 1
N DATA1 BITS
TPC
N TPC BITS
DATA 2
N DATA2 BITS
PILOT
N PILOT BITS
T SLOT = 0.625ms. 10*2 K BITS (k=0..7)

FIG. 2A

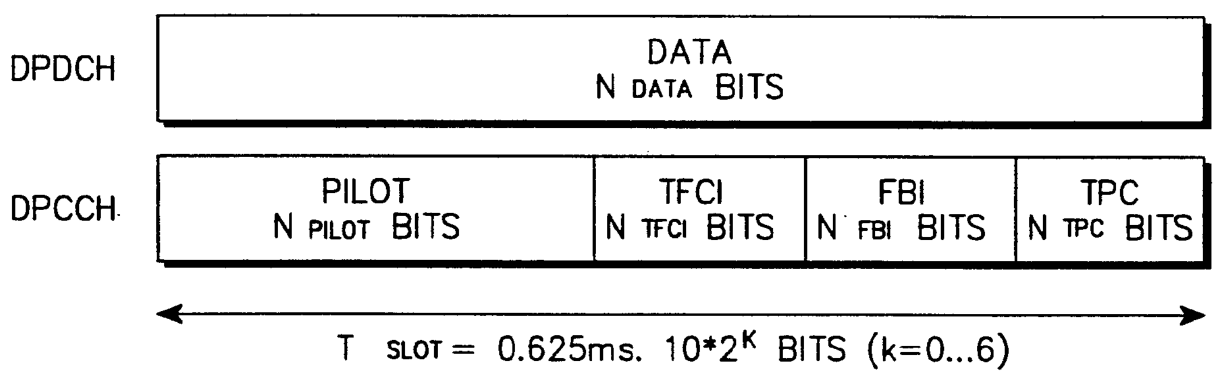

DPDCH
DATA
N DATA BITS
DPCCH
PILOT
N PILOT BITS
TFCI
N TFCI BITS
FBI
N FBI BITS
TPC
N TPC BITS
T SLOT = 0.625ms. 10*2K BITS (k=0...6)

FIG. 2B

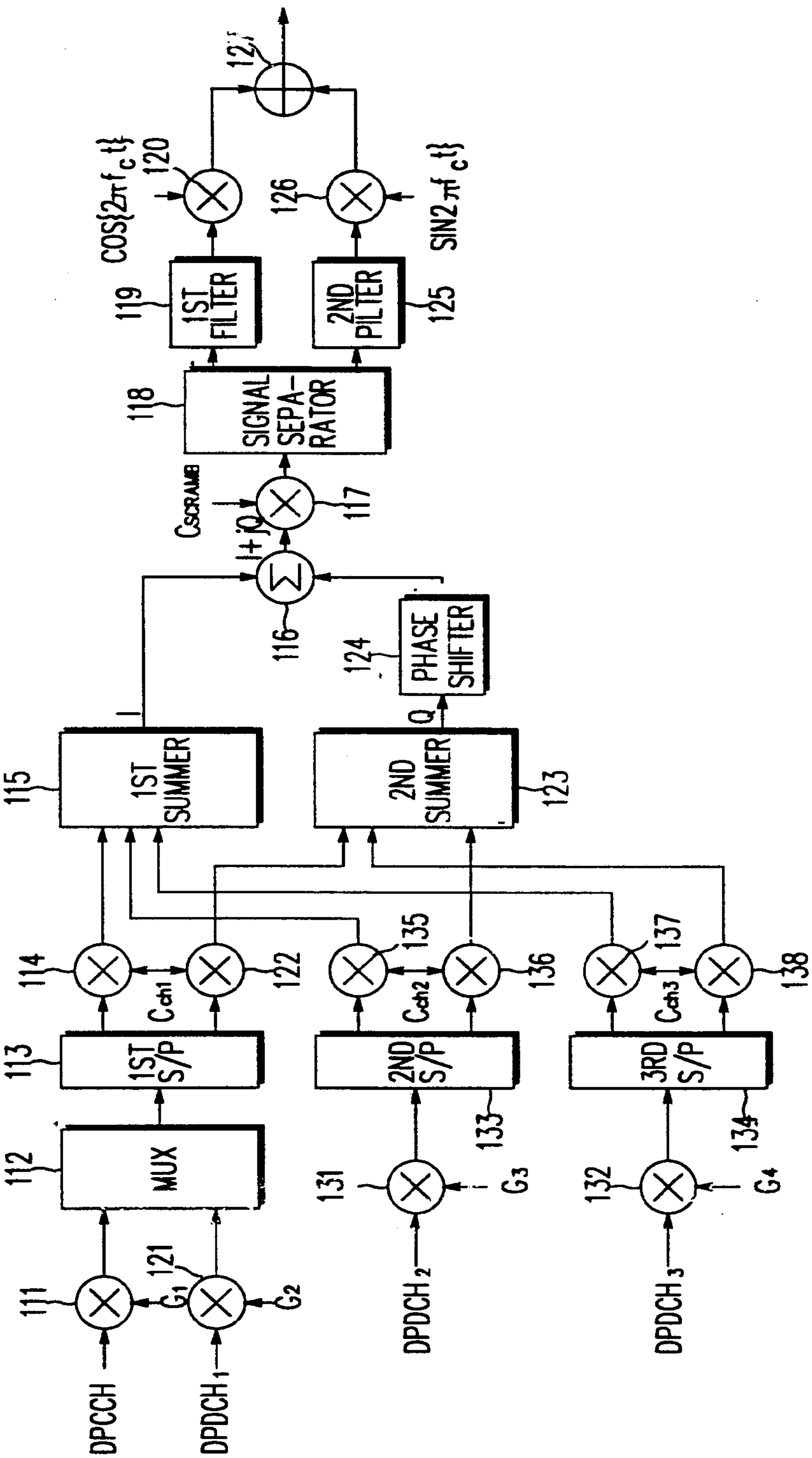

DPCCH
DPDCH1
DPDCH2
DPDCH3
G1
G2
G3
G4
111
121
131
132
112
MUX
113
1ST S/P
133
2ND S/P
134
3RD S/P
Cch1
Cch2
Cch3
114
122
135
136
137
138
115
1ST SUMMER
123
2ND SUMMER
I
Q
124
PHASE SHIFTER
116
I+jQ
CSCRAMB
117
118
SIGNAL SEPA-RATOR
119
1ST FILTER
125
2ND PILTER
120
126
COS{2πfct}
SIN2πfct}

FIG. 3A

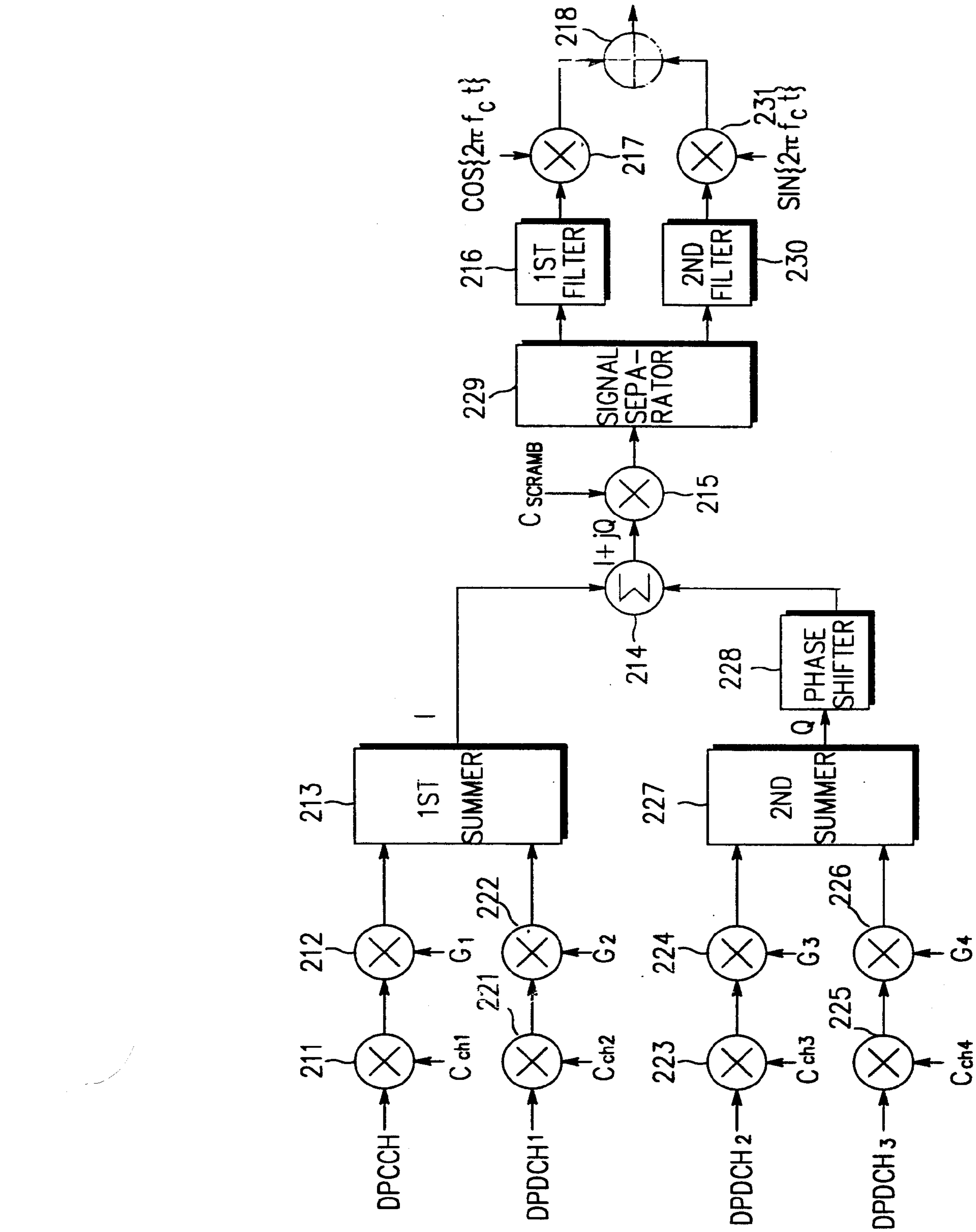

DPCCH
DPDCH1
DPDCH2
DPDCH3
211
212
221
222
223
224
225
226
Cch1
Cch2
Cch3
Cch4
G1
G2
G3
G4
213
1ST SUMMER
227
2ND SUMMER
I
Q
228
PHASE SHIFTER
214
I+jQ
C SCRAMB
215
229
SIGNAL SEPA-RATOR
216
1ST FILTER
230
2ND FILTER
COS{2π fc t}
SIN{2π fc t}
217
231
218

FIG. 3B

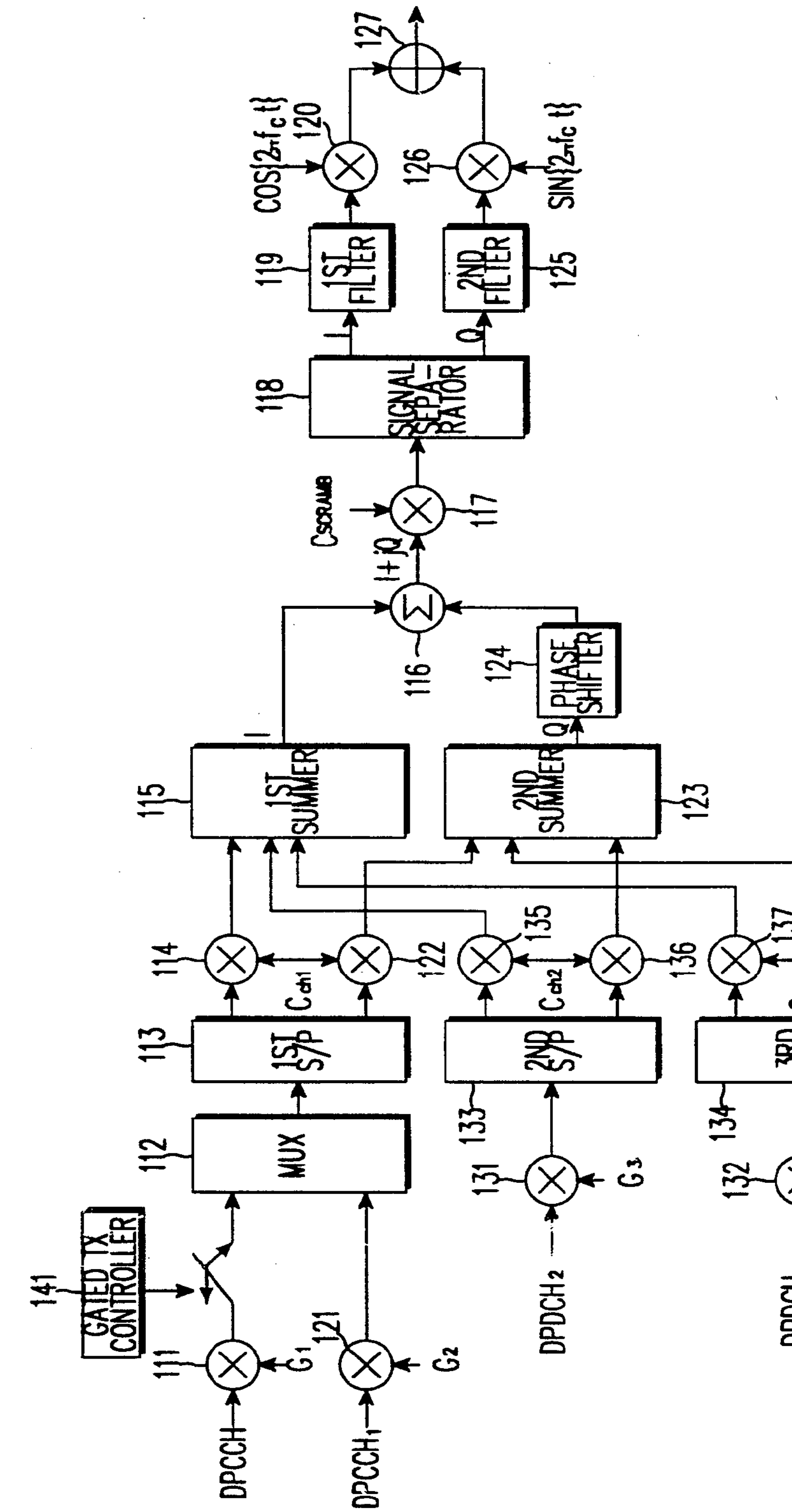

DPCCH
DPCCH1
DPDCH2
DPDCH3
G1
G2
G3
G4
111
121
131
132
141
GATED TX CONTROLLER
112
MUX
113
1ST S/P
133
2ND S/P
134
3RD S/P
Cch1
Cch2
Cch3
114
122
135
136
137
138
115
1ST SUMMER
123
2ND SUMMER
I
Q
124
PHASE SHIFTER
116
I+jQ
CSCRAMB
117
118
SIGNAL SEPA-RATOR
119
1ST FILTER
125
2ND FILTER
COS{2πfc t}
SIN{2πfc t}
120
126
127

FIG. 4A

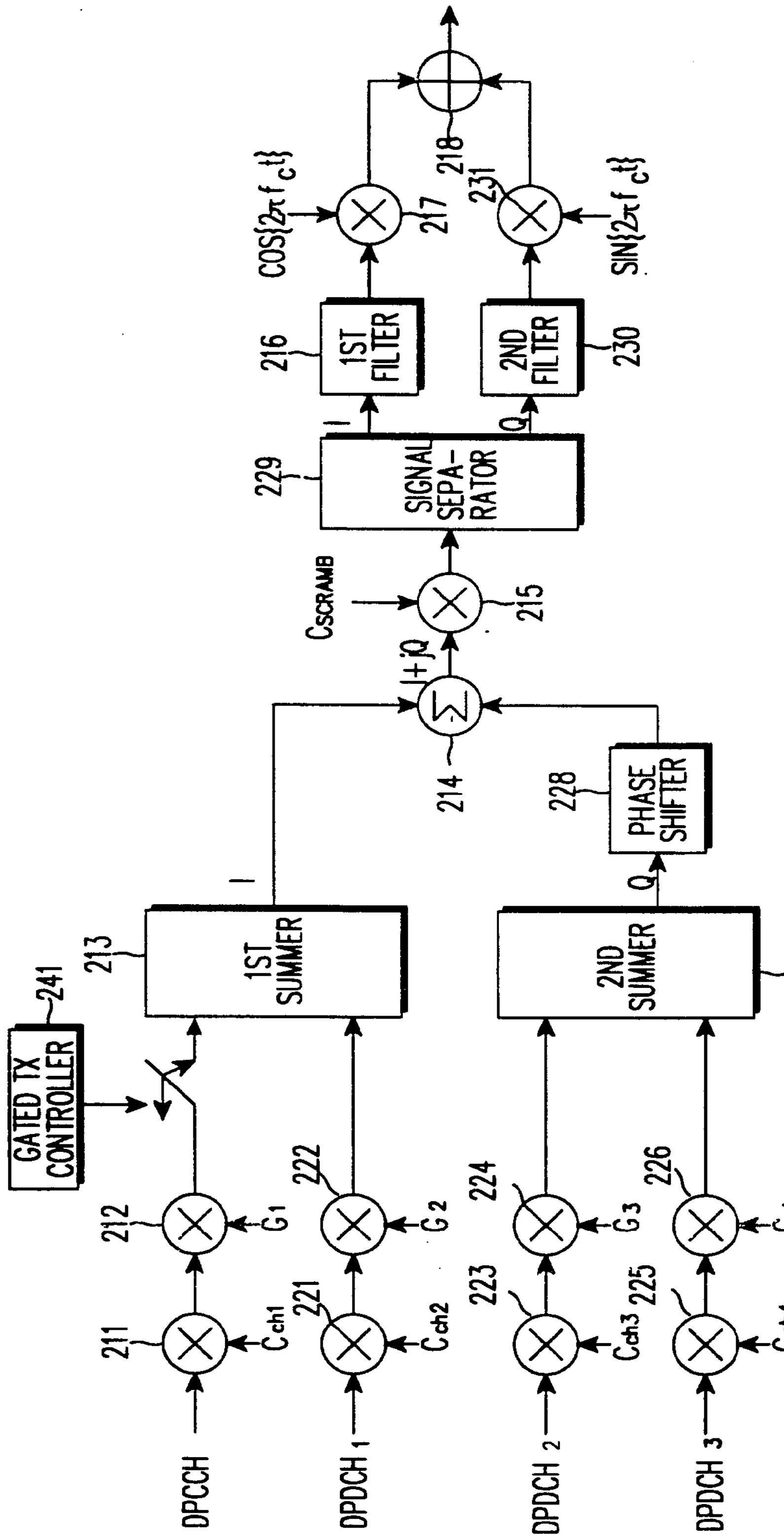

DPCCH
DPDCH 1
DPDCH 2
DPDCH 3
211
221
223
225
Cch1
Cch2
Cch3
Cch4
212
222
224
226
G1
G2
G3
G4
GATED TX CONTROLLER
241
213
1ST SUMMER
227
2ND SUMMER
I
Q
228
PHASE SHIFTER
214
I+jQ
CSCRAMB
215
229
SIGNAL SEPA-RATOR
I
Q
216
1ST FILTER
230
2ND FILTER
COS{2πfct}
SIN{2πfct}
217
231
218

FIG. 4B

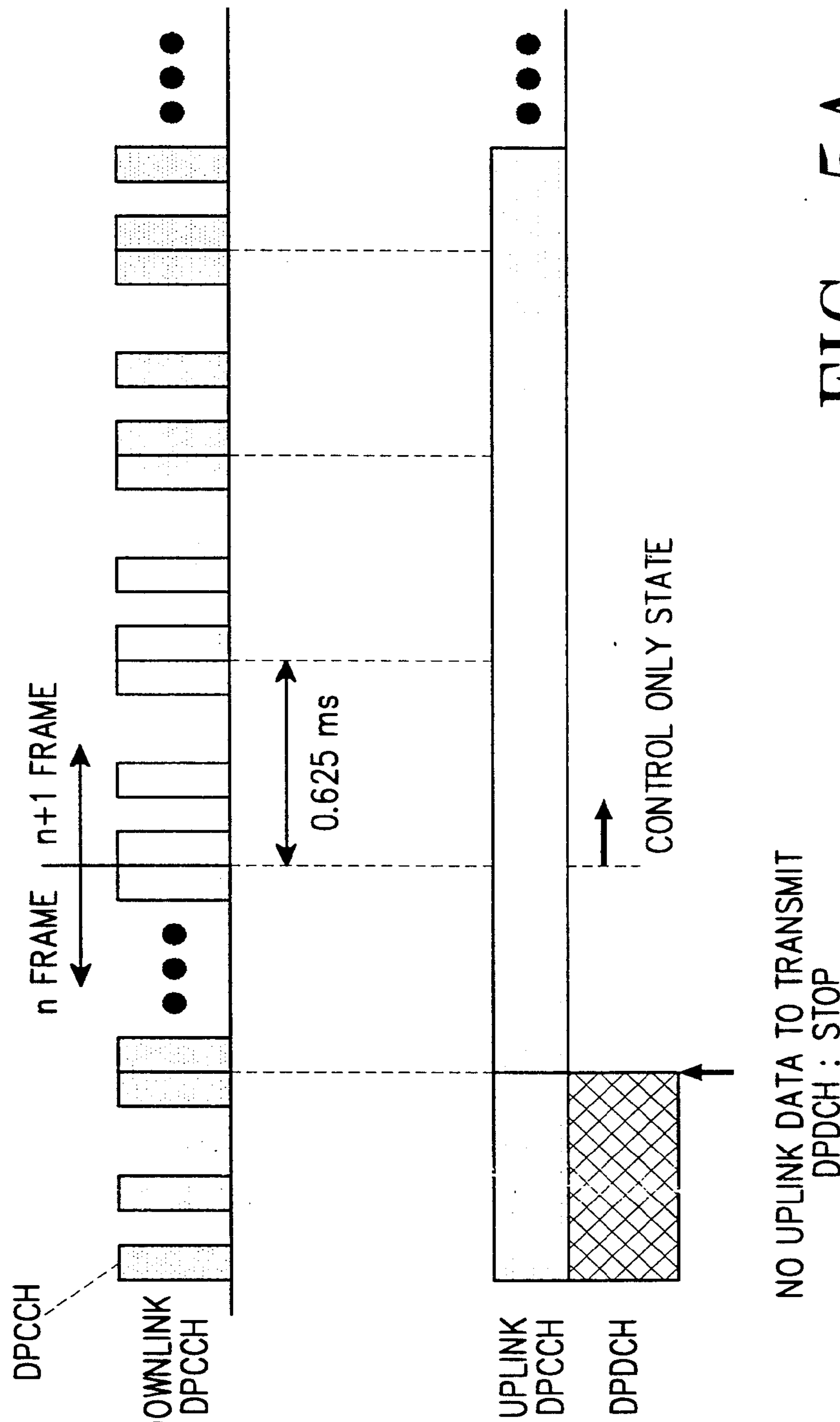

DPCCH
DOWNLINK DPCCH
n FRAME
n+1 FRAME
0.625 ms
UPLINK DPCCH
DPDCH
CONTROL ONLY STATE
NO UPLINK DATA TO TRANSMIT
DPDCH : STOP

FIG. 5A

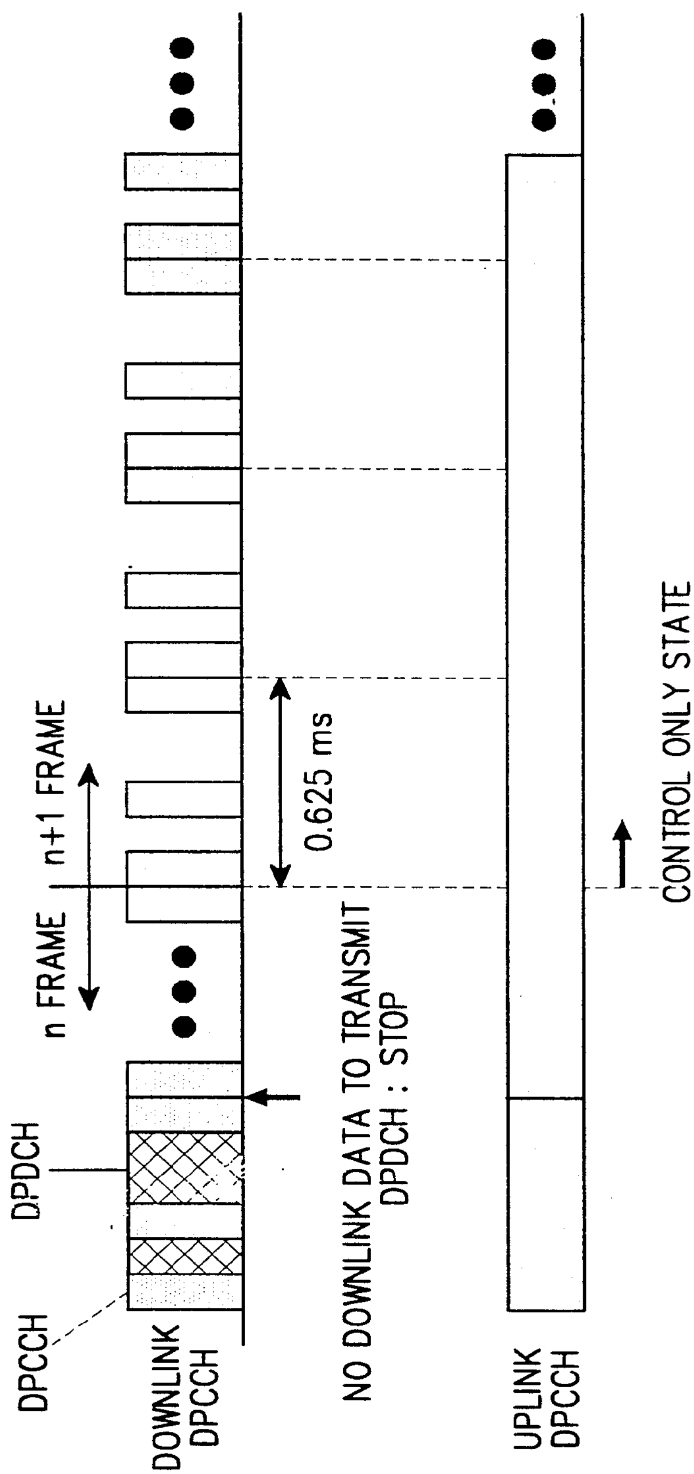
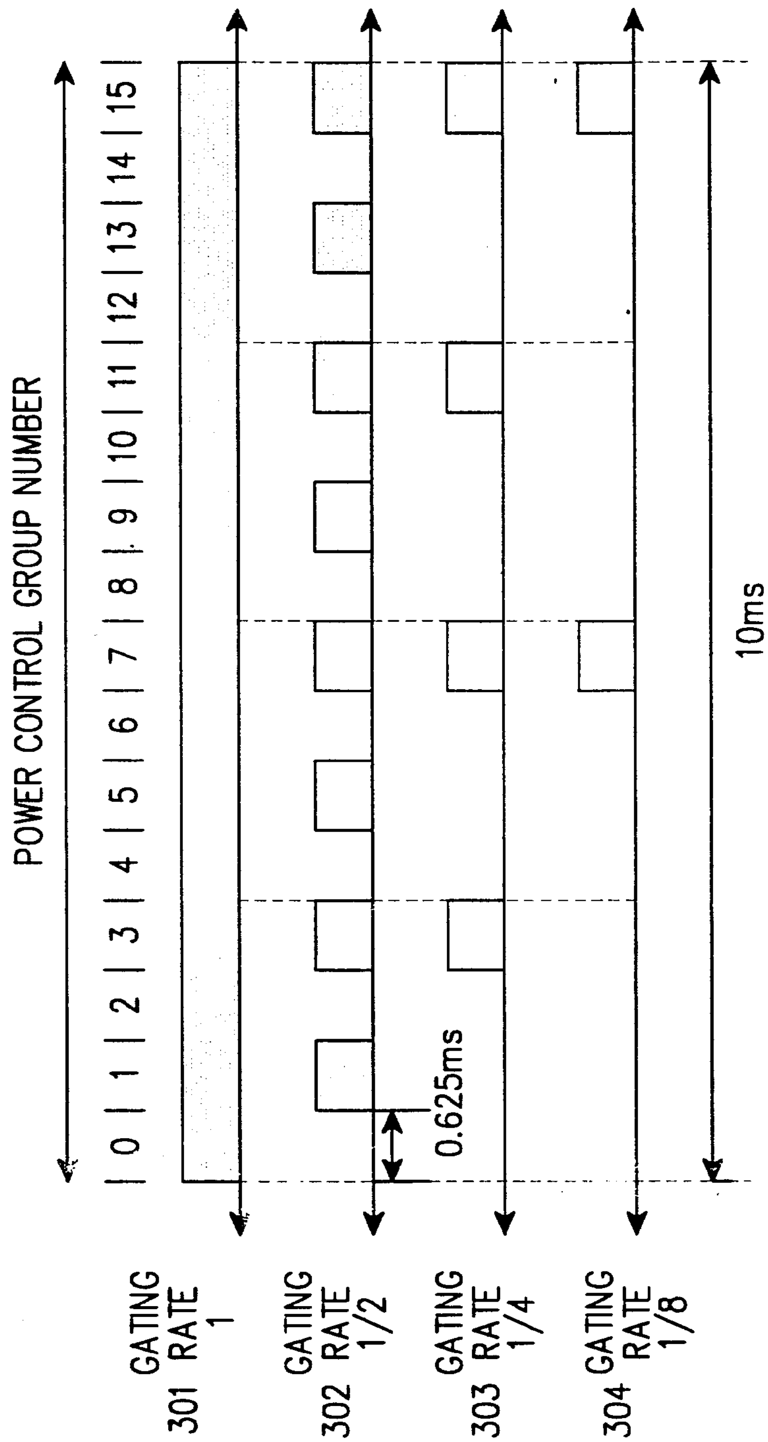

DPCCH
DPDCH
DOWNLINK DPCCH
n FRAME
n+1 FRAME
0.625 ms
NO DOWNLINK DATA TO TRANSMIT
DPDCH : STOP
UPLINK DPCCH
CONTROL ONLY STATE

FIG. 5B

FIG. 6A


POWER CONTROL GROUP NUMBER
0 | 1 | 2 | 3 | 4 | 5 | 6 | 7 | 8 | 9 | 10 | 11 | 12 | 13 | 14 | 15
0.625ms
10ms
301 GATING RATE 1
302 GATING RATE 1/2
303 GATING RATE 1/4
304 GATING RATE 1/8

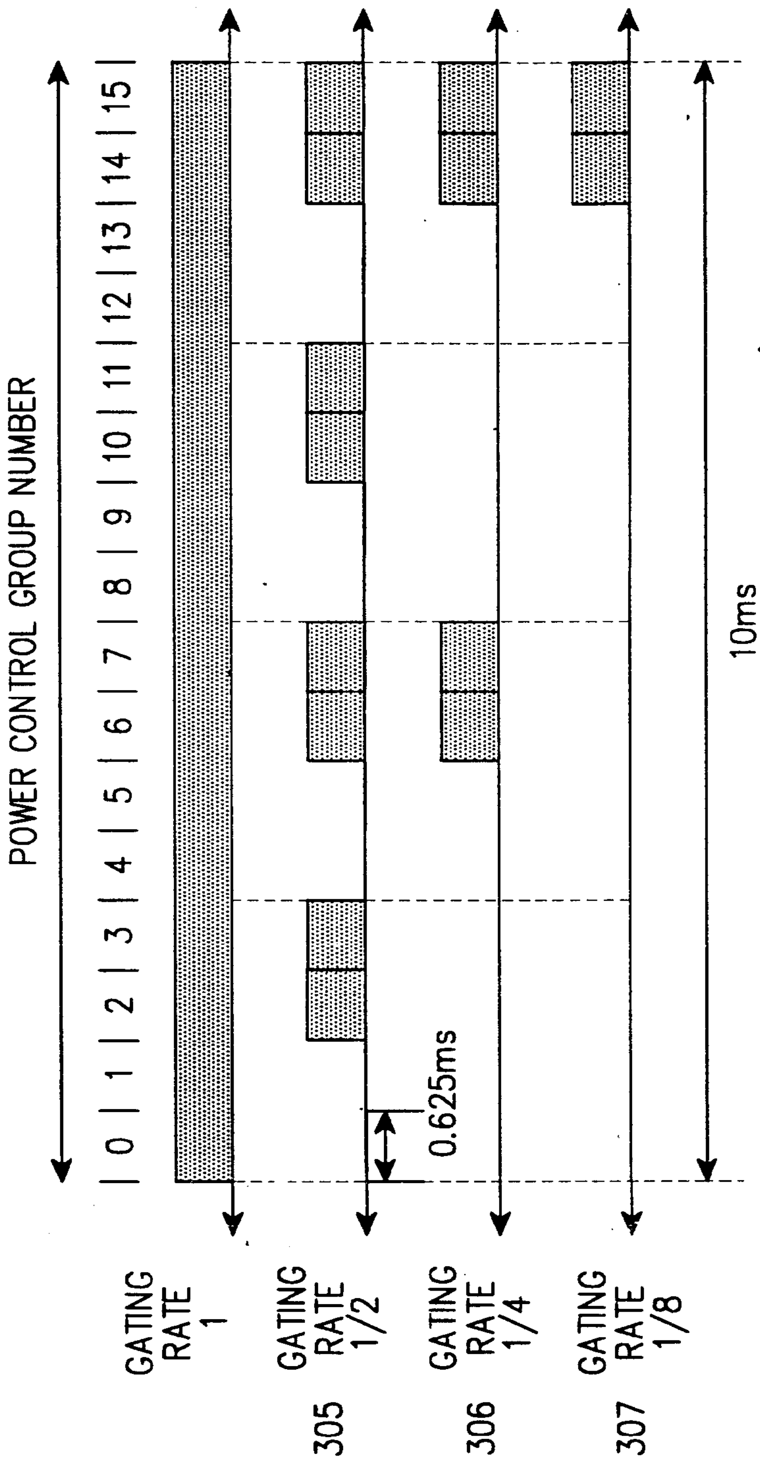

POWER CONTROL GROUP NUMBER
0
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
0.625ms
10ms
GATING RATE 1
305 GATING RATE 1/2
306 GATING RATE 1/4
307 GATING RATE 1/8

FIG. 6B

FIG. 7A

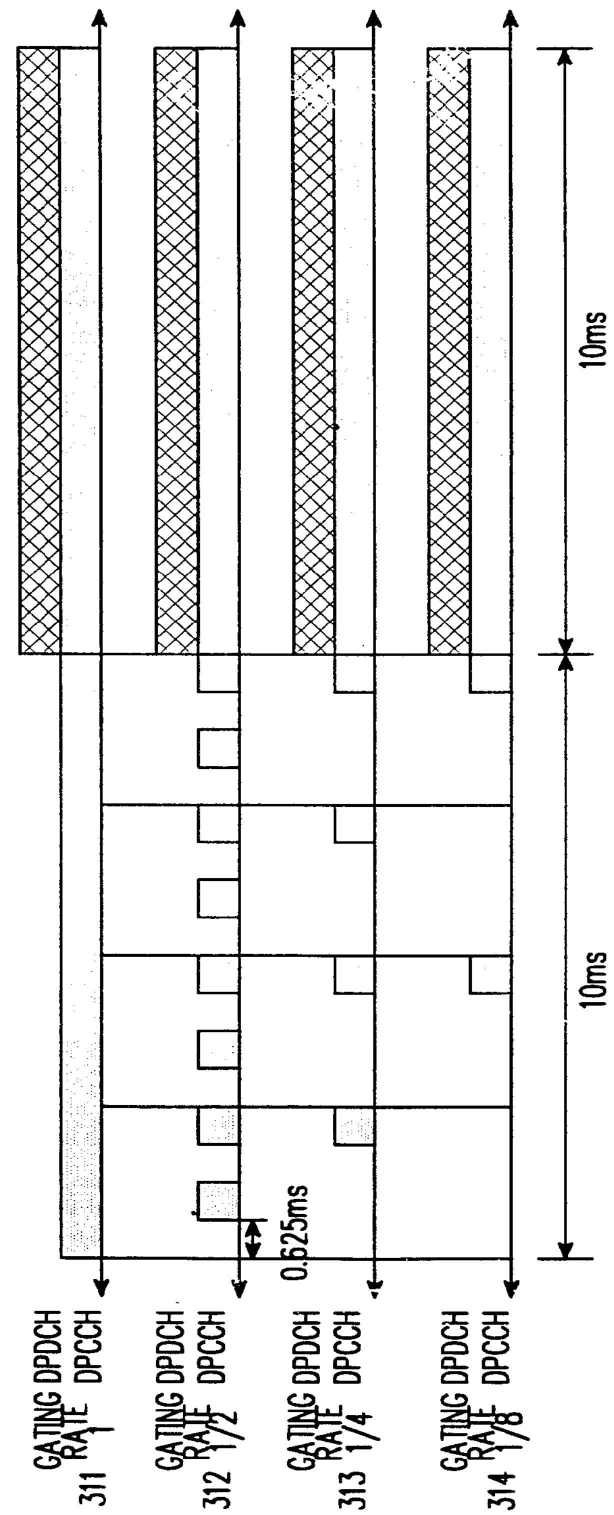
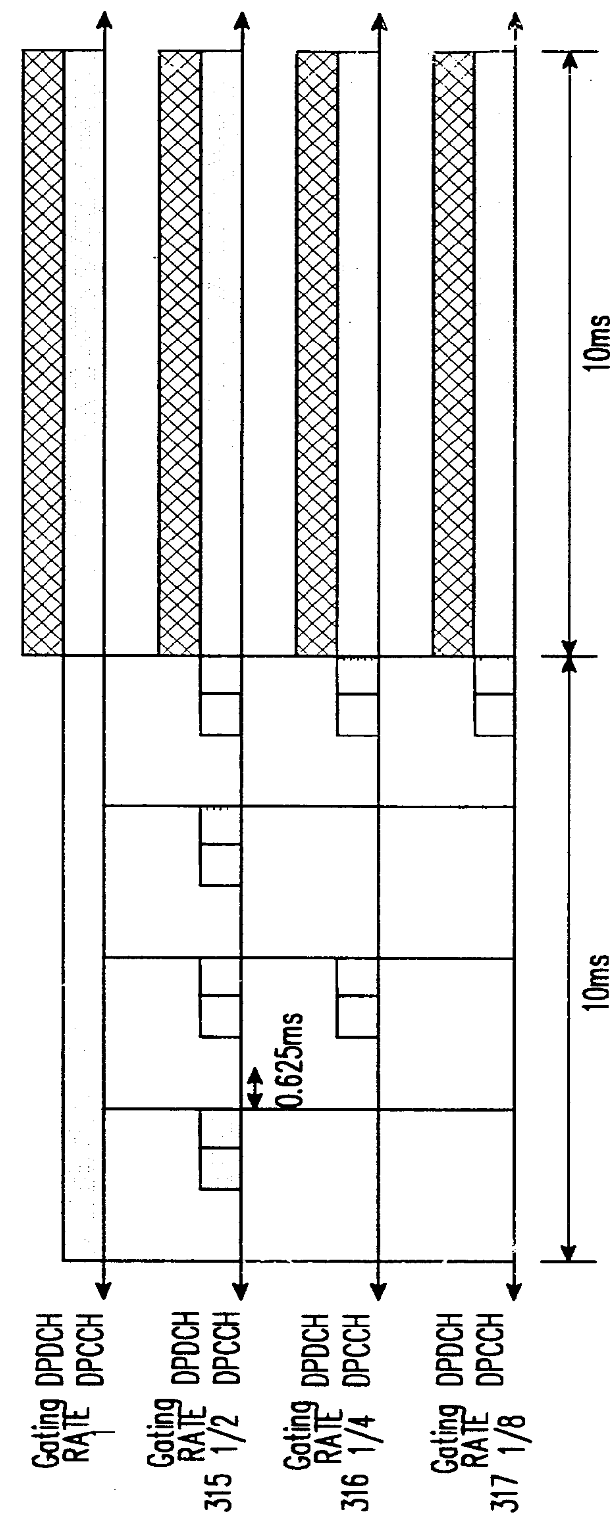

311 GATING RATE 1
DPDCH
DPCCH
312 GATING RATE 1/2
DPDCH
DPCCH
313 GATING RATE 1/4
DPDCH
DPCCH
314 GATING RATE 1/8
DPDCH
DPCCH
0.625ms
10ms
10ms

FIG. 7B


Gating RATE 1 DPDCH DPCCH
315 Gating RATE 1/2 DPDCH DPCCH
316 Gating RATE 1/4 DPDCH DPCCH
317 Gating RATE 1/8 DPDCH DPCCH
0.625ms
10ms
10ms

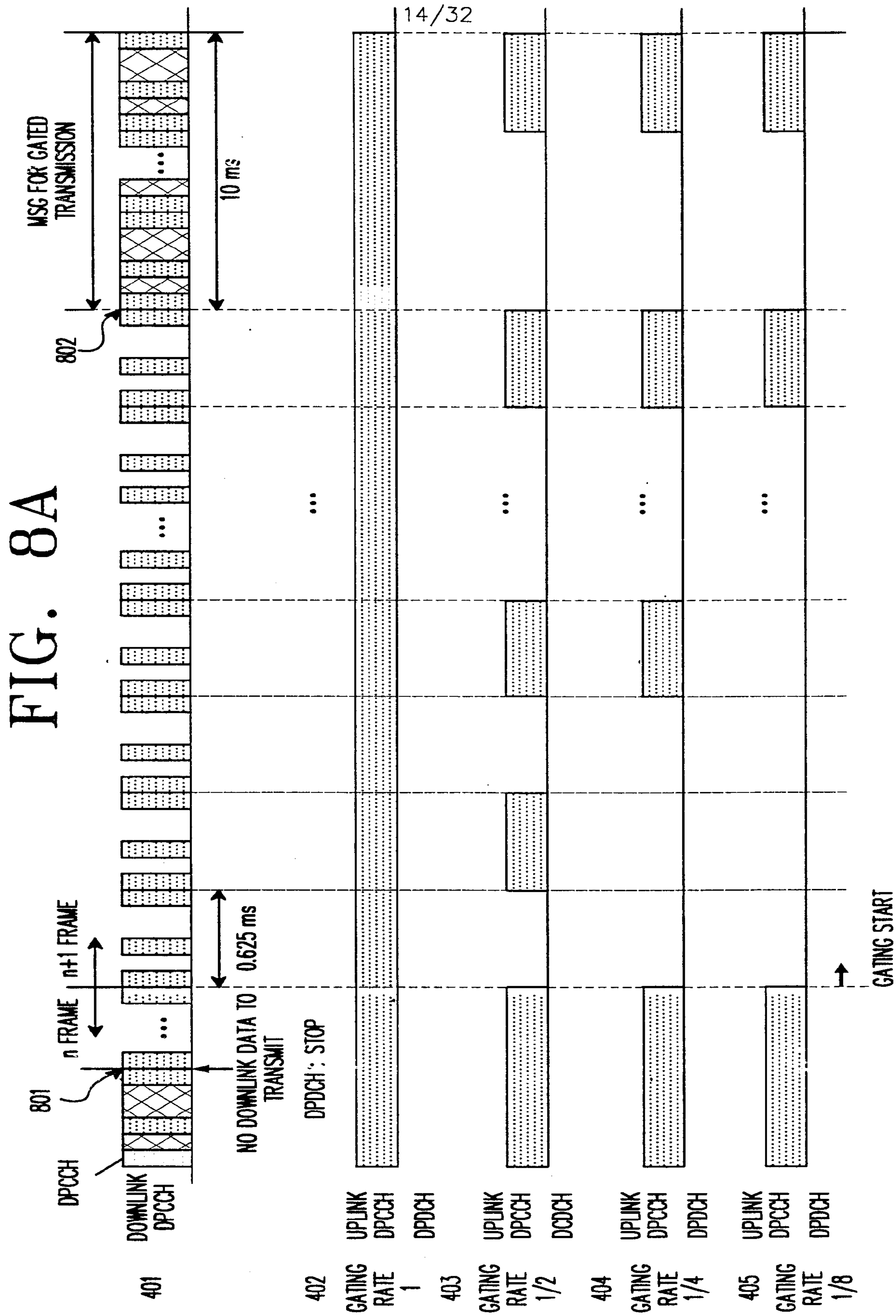
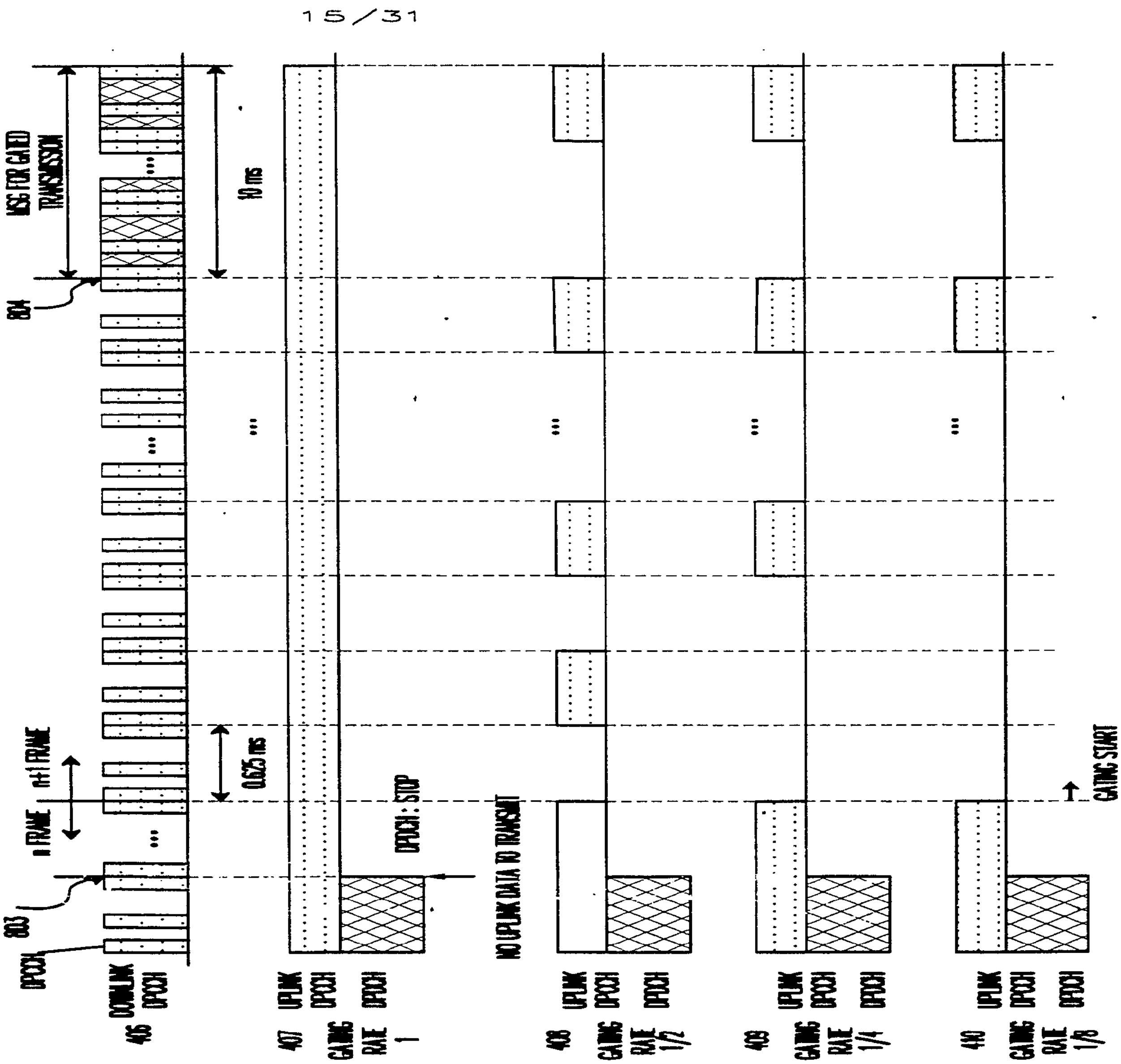

MSG FOR GATED TRANSMISSION
10 ms
802
801
DPCCH
n FRAME
n+1 FRAME
0.625 ms
NO DOWNLINK DATA TO TRANSMIT
DPDCH : STOP
GATING START
401
DOWNLINK DPCCH
402
GATING RATE 1
UPLINK DPCCH
DPDCH
403
GATING RATE 1/2
UPLINK DPCCH
DCDCH
404
GATING RATE 1/4
UPLINK DPCCH
DPDCH
405
GATING RATE 1/8
UPLINK DPCCH
DPDCH

FIG. 8A

FIG. 8B


15/31
DPCCH
803
804
406 DOWNLINK DPCCH
MSG FOR GATED TRANSMISSION
10 ms
n FRAME
n+1 FRAME
0.625 ms
DPCCH : STOP
NO UPLINK DATA TO TRANSMIT
407 GATING RATE 1
UPLINK DPCCH DPDCH
408 GATING RATE 1/2
UPLINK DPCCH DPDCH
409 GATING RATE 1/4
UPLINK DPCCH DPDCH
410 GATING RATE 1/8
UPLINK DPCCH DPDCH
GATING START

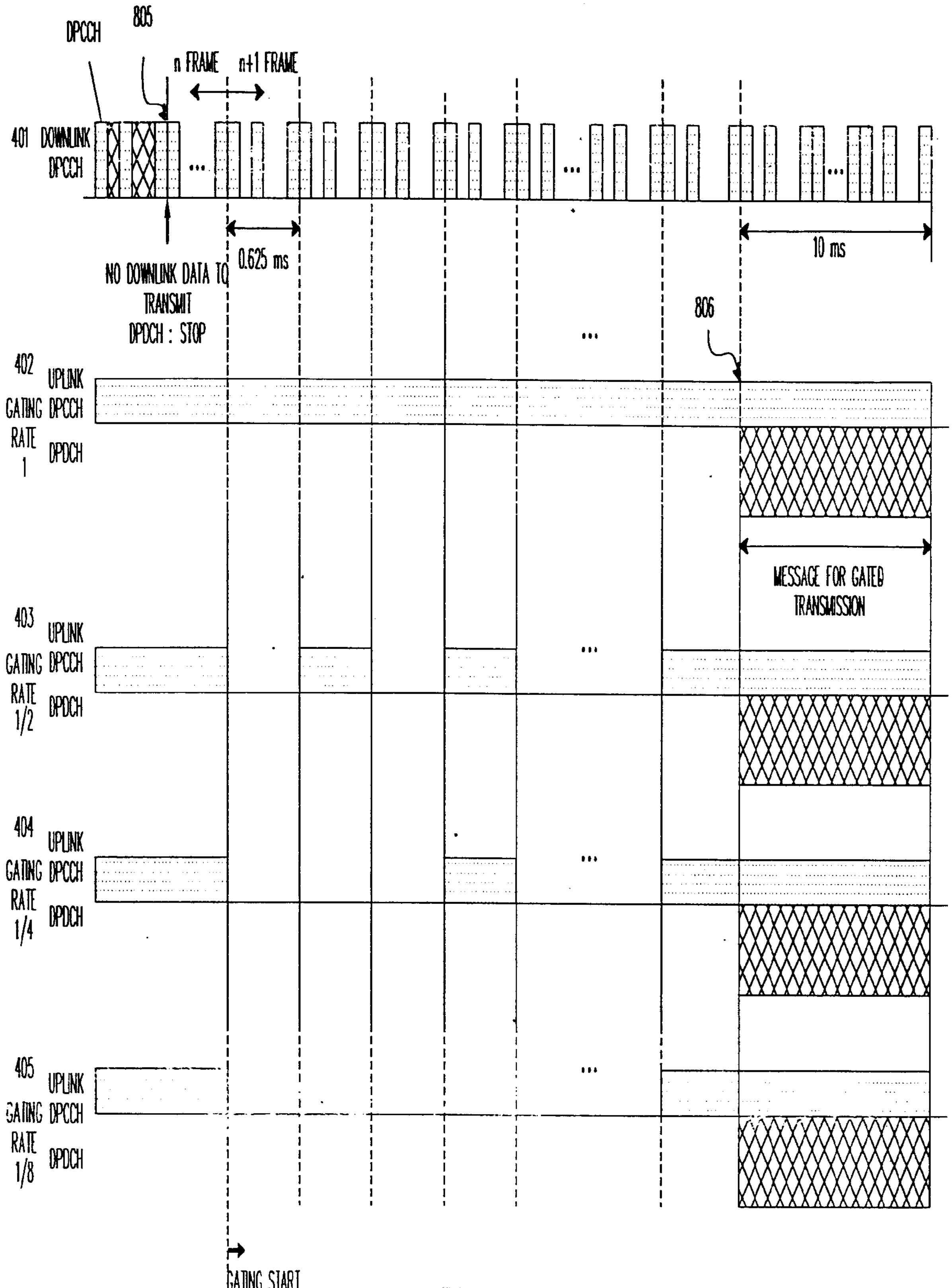
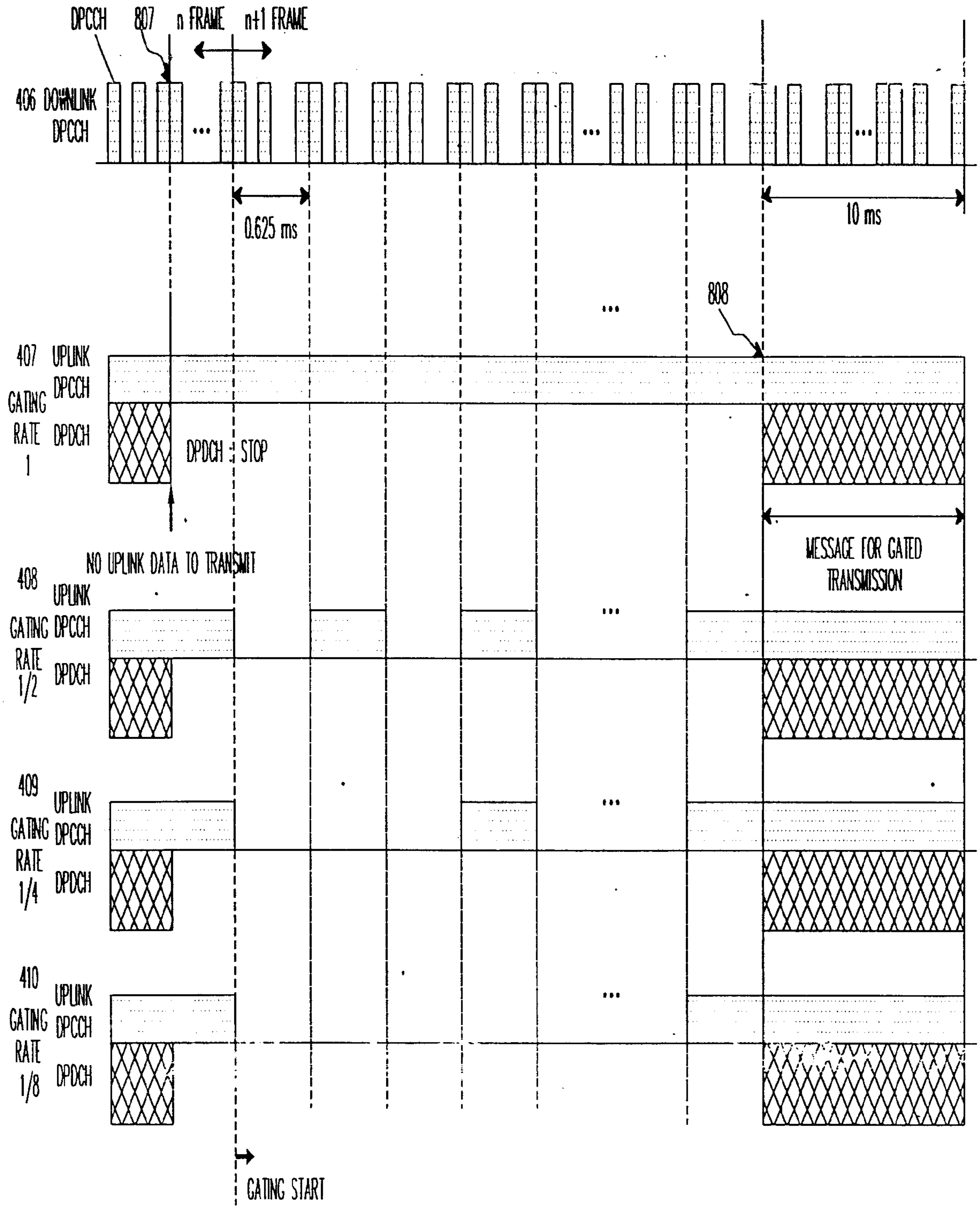

DPCCH
805
n FRAME
n+1 FRAME
401 DOWNLINK DPCCH
0.625 ms
10 ms
NO DOWNLINK DATA TO TRANSMIT
DPDCH : STOP
806
402 UPLINK GATING RATE 1
DPCCH
DPDCH
MESSAGE FOR GATED TRANSMISSION
403 UPLINK GATING RATE 1/2
DPCCH
DPDCH
404 UPLINK GATING RATE 1/4
DPCCH
DPDCH
405 UPLINK GATING RATE 1/8
DPCCH
DPDCH
GATING START

FIG. 8C



DPCCH
807
n FRAME
n+1 FRAME
406 DOWNLINK DPCCH
0.625 ms
10 ms
808
407 UPLINK GATING RATE 1
DPCCH
DPDCH
DPDCH : STOP
NO UPLINK DATA TO TRANSMIT
MESSAGE FOR GATED TRANSMISSION
408 UPLINK GATING RATE 1/2
409 UPLINK GATING RATE 1/4
410 UPLINK GATING RATE 1/8
GATING START

FIG. 8D

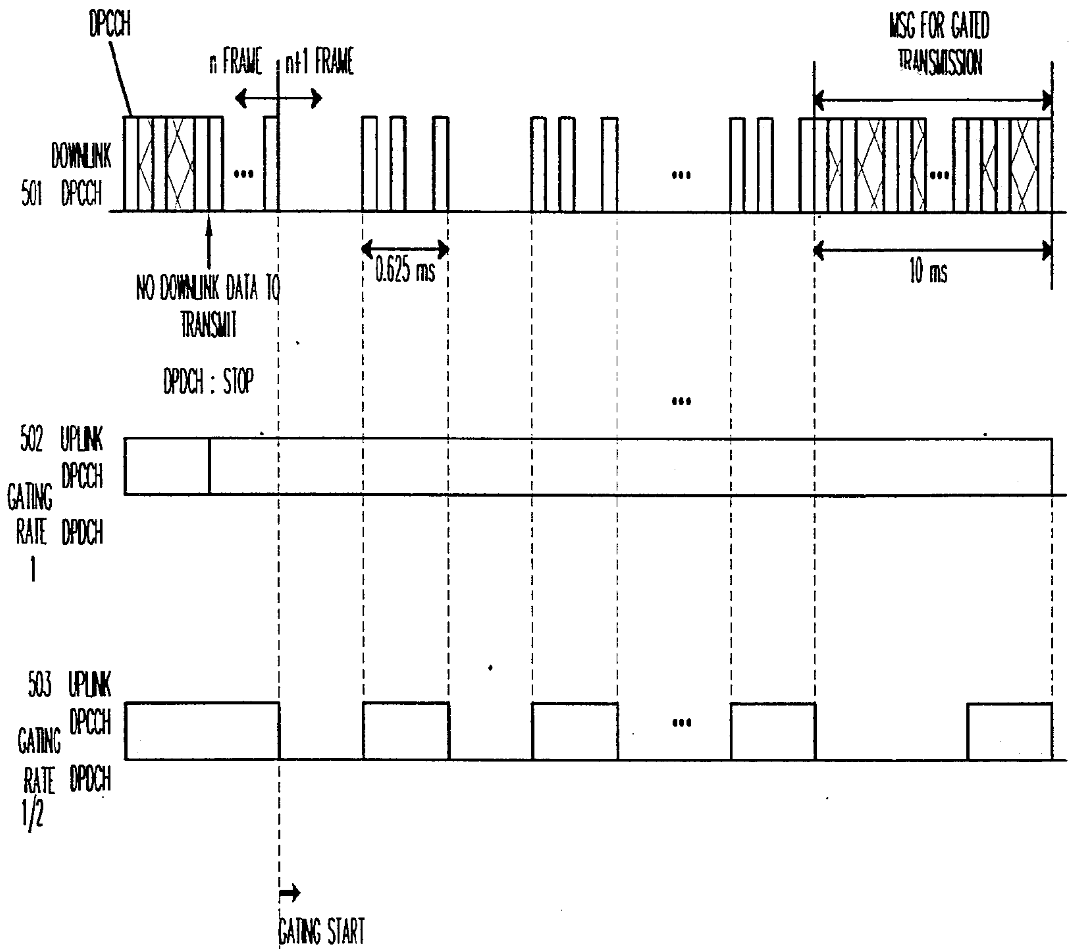

DPCCH
n FRAME
n+1 FRAME
MSG FOR GATED TRANSMISSION
DOWNLINK
501 DPCCH
0.625 ms
10 ms
NO DOWNLINK DATA TO TRANSMIT
DPDCH : STOP
502 UPLINK DPCCH
GATING RATE 1 DPDCH
503 UPLINK DPCCH
GATING RATE 1/2 DPDCH
GATING START

FIG. 9A

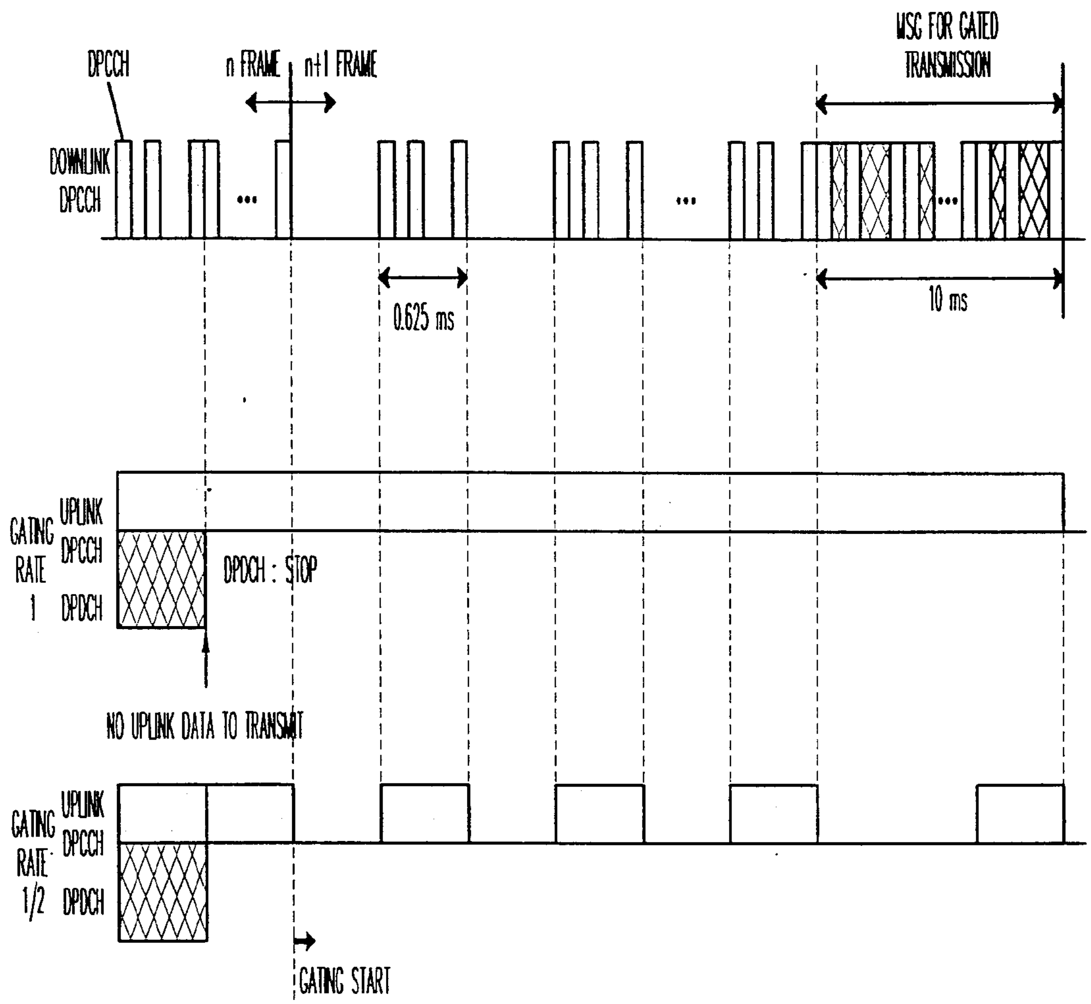

DPCCH
n FRAME
n+1 FRAME
MSC FOR GATED TRANSMISSION
DOWNLINK DPCCH
0.625 ms
10 ms
GATING RATE 1
UPLINK DPCCH
DPDCH
DPDCH : STOP
NO UPLINK DATA TO TRANSMIT
GATING RATE 1/2
UPLINK DPCCH
DPDCH
GATING START

FIG. 9B

FIG. 10A

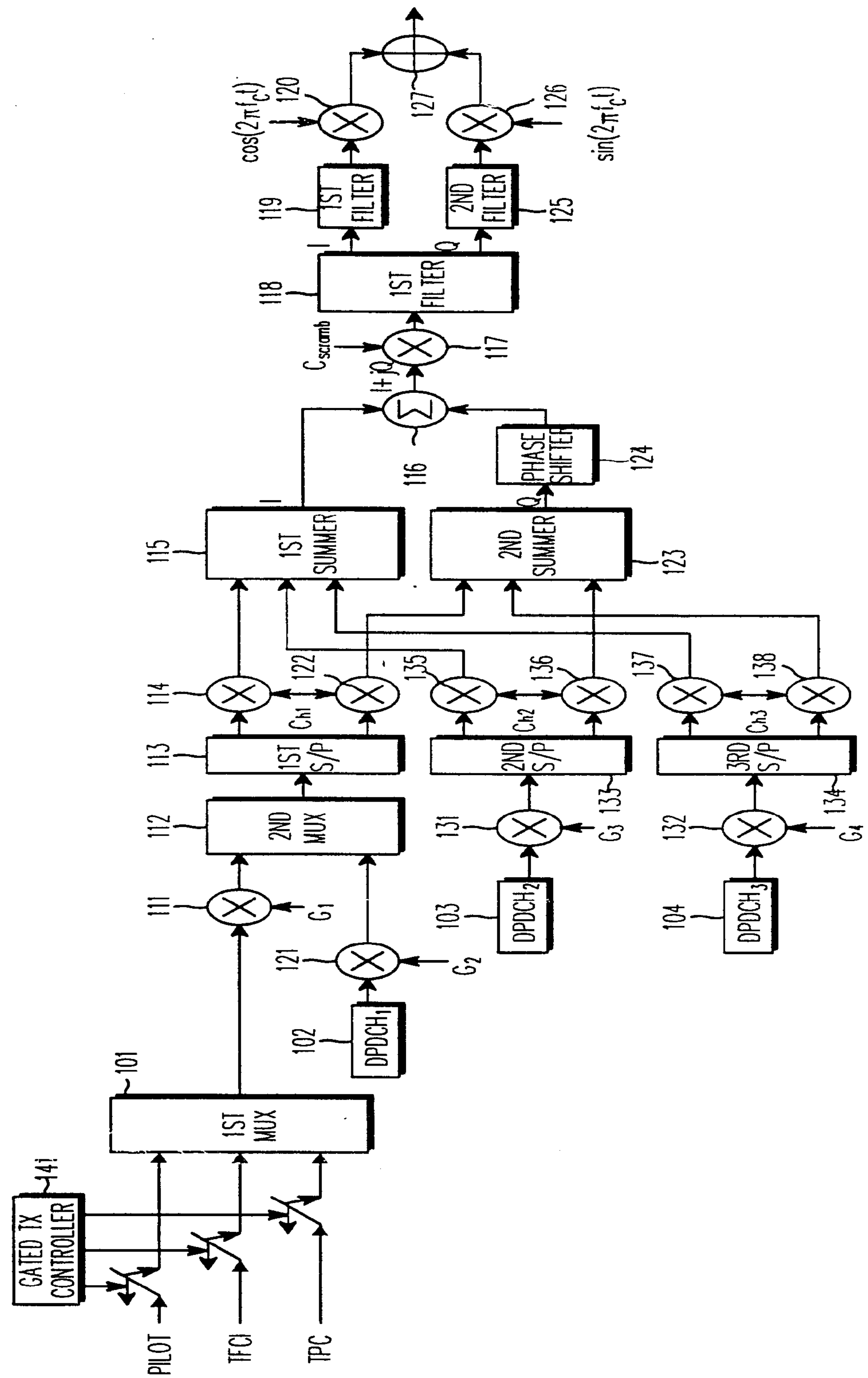
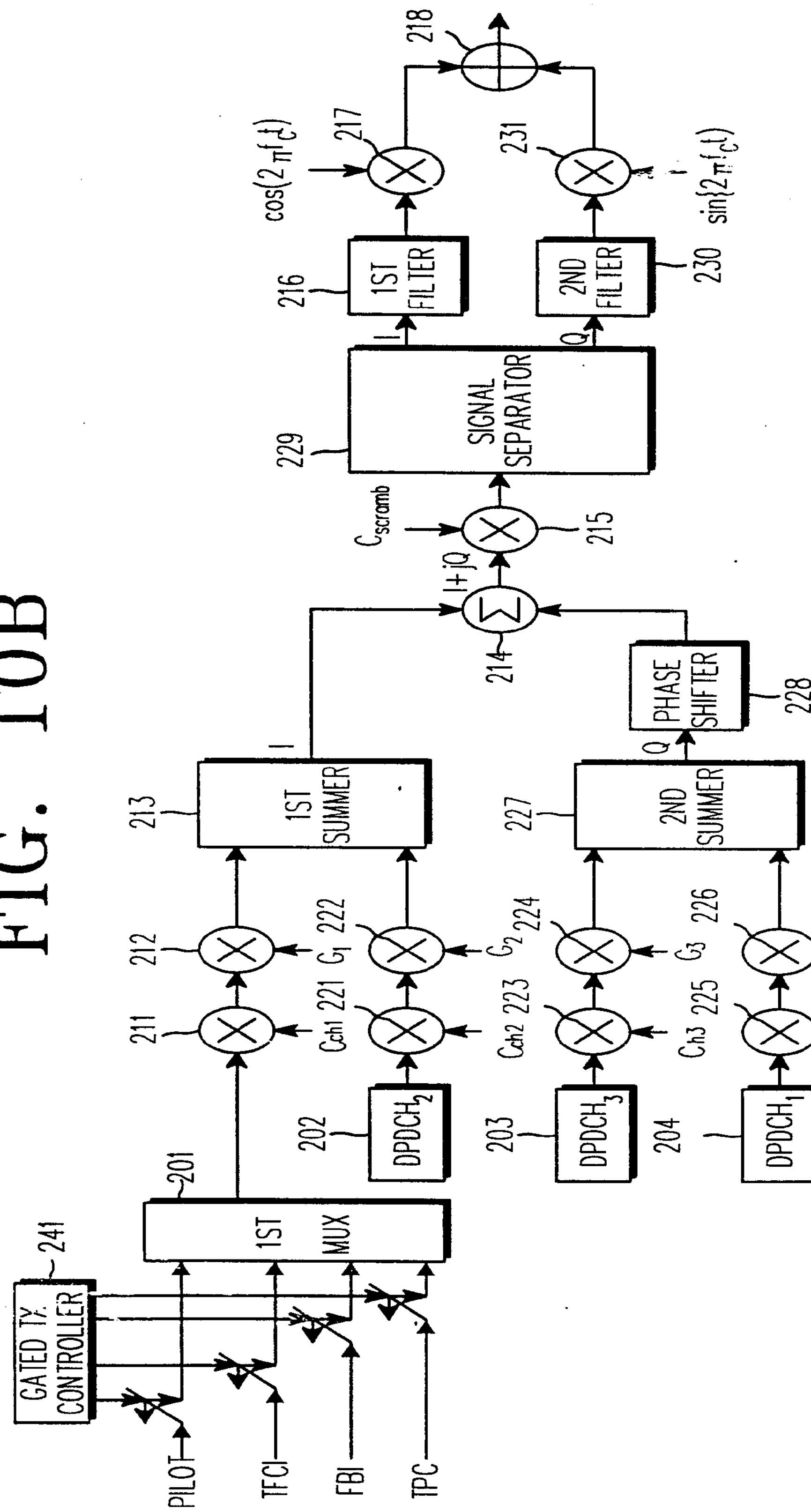

GATED TX CONTROLLER
141
PILOT
TFCI
TPC
1ST MUX
101
DPDCH1
102
121
G2
111
G1
2ND MUX
112
1ST S/P
113
114
122
Ch1
DPDCH2
103
131
G3
2ND S/P
133
135
136
Ch2
DPDCH3
104
132
G4
3RD S/P
134
137
138
Ch3
1ST SUMMER
115
2ND SUMMER
123
I
Q
PHASE SHIFTER
124
116
I+jQ
Cscramb
117
1ST FILTER
118
1ST FILTER
119
2ND FILTER
125
cos(2πfct)
120
sin(2πfct)
126
127

FIG. 10B


GATED TX CONTROLLER
241
1ST MUX
201
PILOT
TFCI
FBI
TPC
DPDCH2
DPDCH3
DPDCH1
202
203
204
211
212
221
222
223
224
225
226
Cch1
Cch2
Ch3
Cch4
G1
G2
G3
G4
1ST SUMMER
213
2ND SUMMER
227
PHASE SHIFTER
228
I
Q
214
I+jQ
Cscramb
215
SIGNAL SEPARATOR
229
1ST FILTER
216
2ND FILTER
230
cos(2πfct)
sin(2πfct)
217
231
218

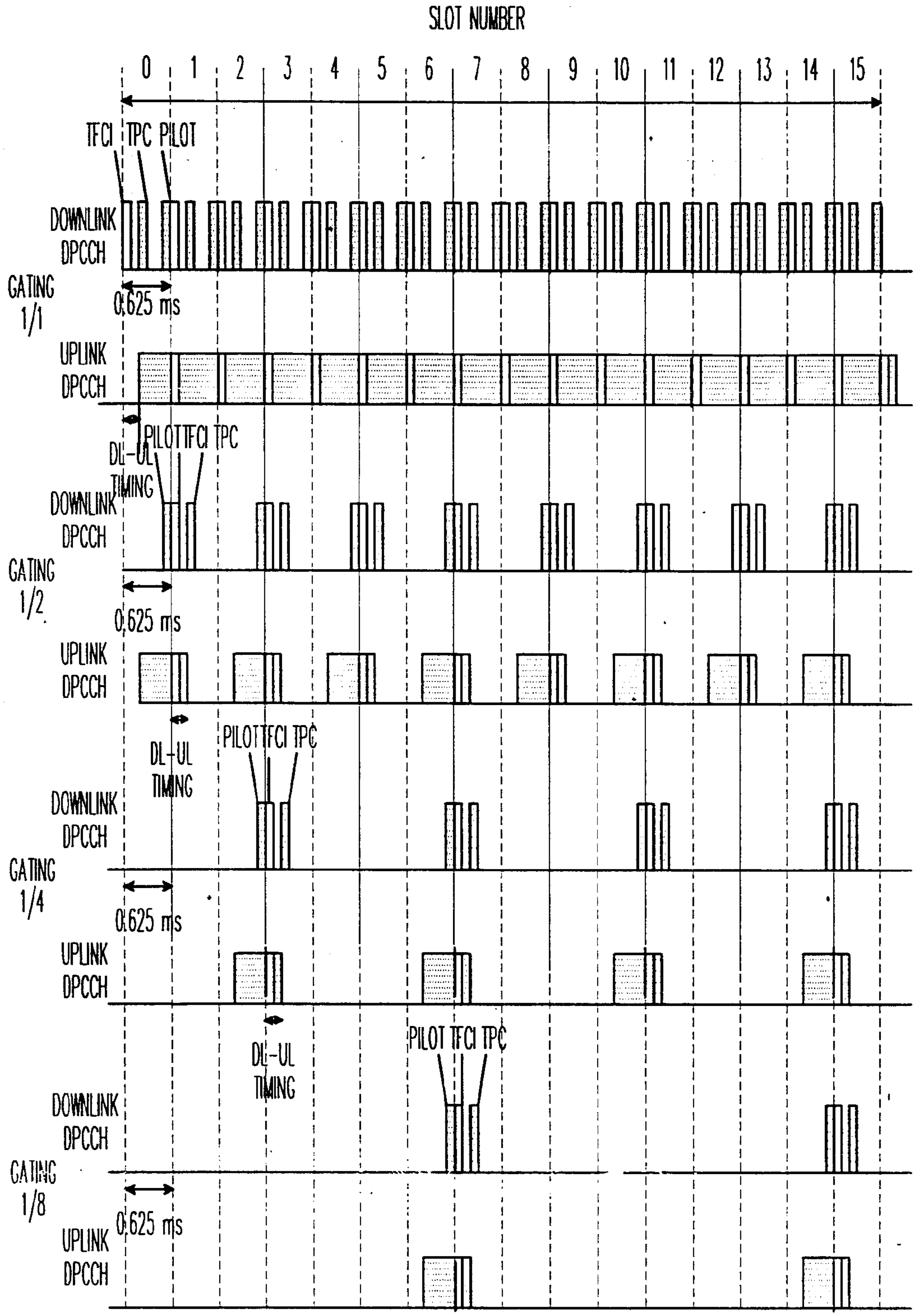

SLOT NUMBER
0 1 2 3 4 5 6 7 8 9 10 11 12 13 14 15
TFCI TPC PILOT
DOWNLINK DPCCH
GATING 1/1
0.625 ms
UPLINK DPCCH
PILOT TFCI TPC
DL-UL TIMING
DOWNLINK DPCCH
GATING 1/2
0.625 ms
UPLINK DPCCH
DL-UL TIMING
PILOT TFCI TPC
DOWNLINK DPCCH
GATING 1/4
0.625 ms
UPLINK DPCCH
DL-UL TIMING
PILOT TFCI TPC
DOWNLINK DPCCH
GATING 1/8
0.625 ms
UPLINK DPCCH

FIG. 11A

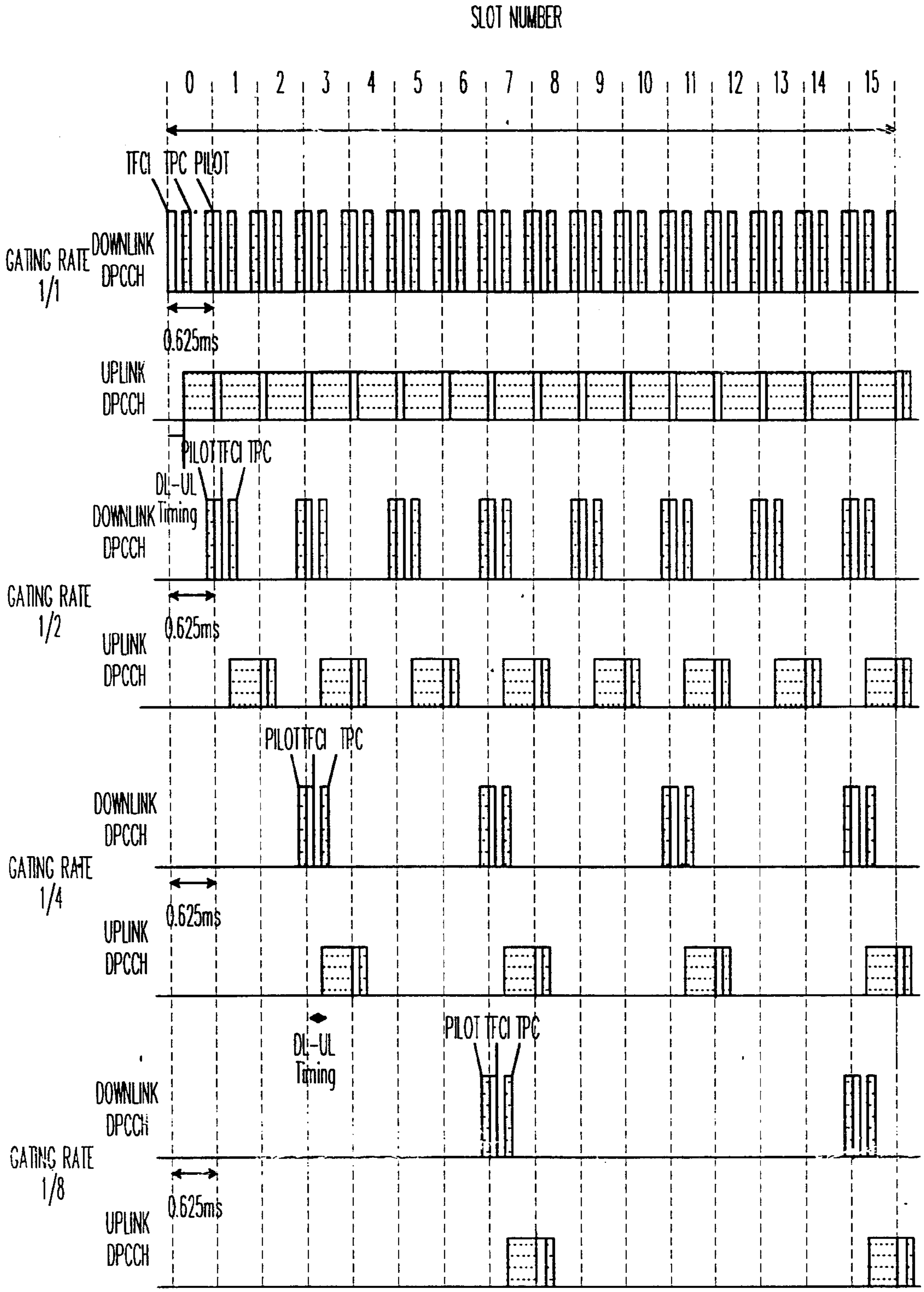

SLOT NUMBER
0
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
TFCI TPC PILOT
GATING RATE 1/1
DOWNLINK DPCCH
0.625ms
UPLINK DPCCH
PILOT TFCI TPC
DL-UL Timing
DOWNLINK DPCCH
GATING RATE 1/2
0.625ms
UPLINK DPCCH
PILOT TFCI TPC
DOWNLINK DPCCH
GATING RATE 1/4
0.625ms
UPLINK DPCCH
DL-UL Timing
PILOT TFCI TPC
DOWNLINK DPCCH
GATING RATE 1/8
0.625ms
UPLINK DPCCH

FIG. 11B

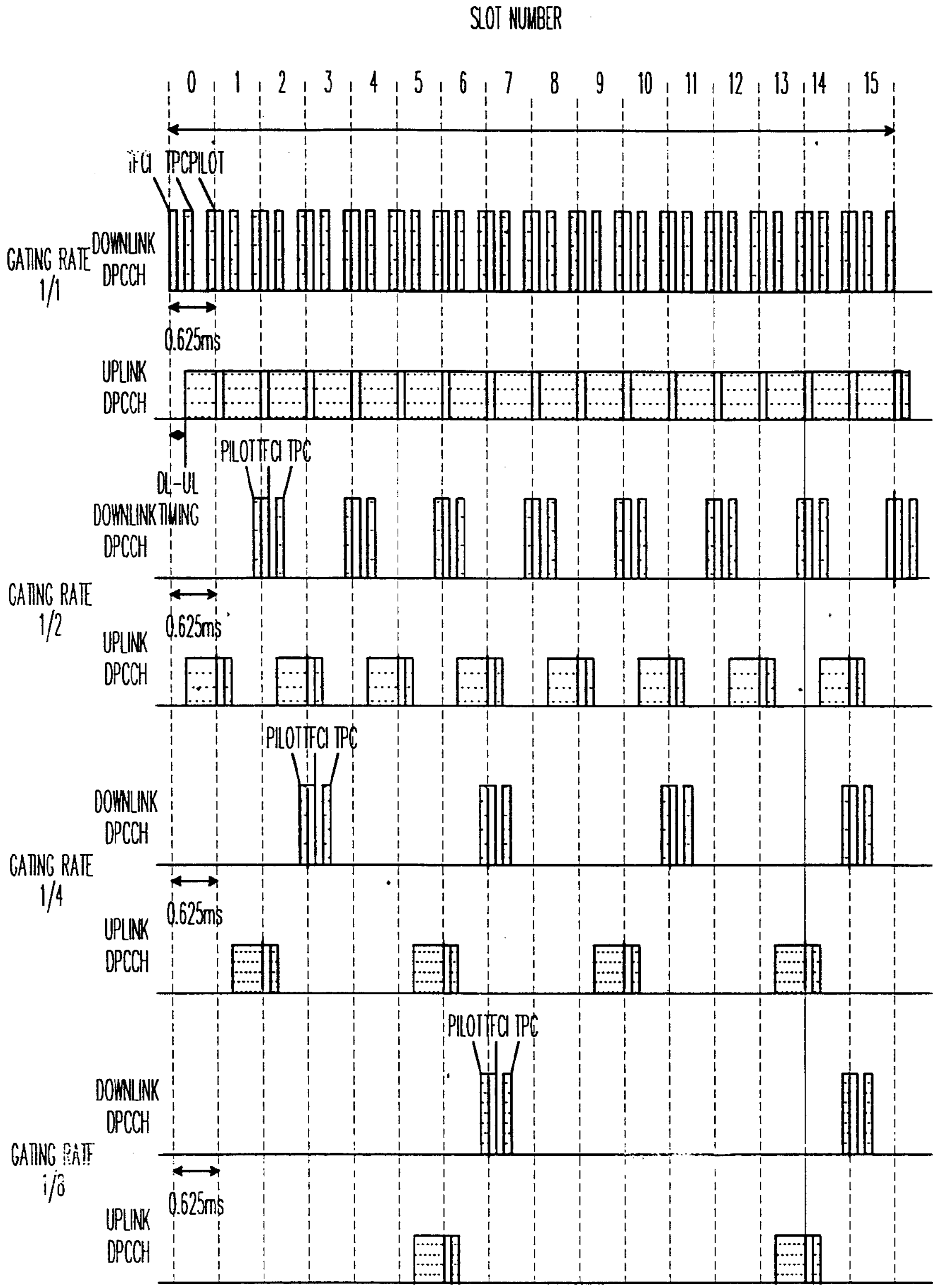

SLOT NUMBER
0
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
TFCI
TPC
PILOT
GATING RATE 1/1
DOWNLINK DPCCH
0.625ms
UPLINK DPCCH
DL-UL TIMING
PILOT
TFCI
TPC
DOWNLINK DPCCH
GATING RATE 1/2
0.625ms
UPLINK DPCCH
PILOT
TFCI
TPC
DOWNLINK DPCCH
GATING RATE 1/4
0.625ms
UPLINK DPCCH
PILOT
TFCI
TPC
DOWNLINK DPCCH
GATING RATE 1/8
0.625ms
UPLINK DPCCH

FIG. 11C

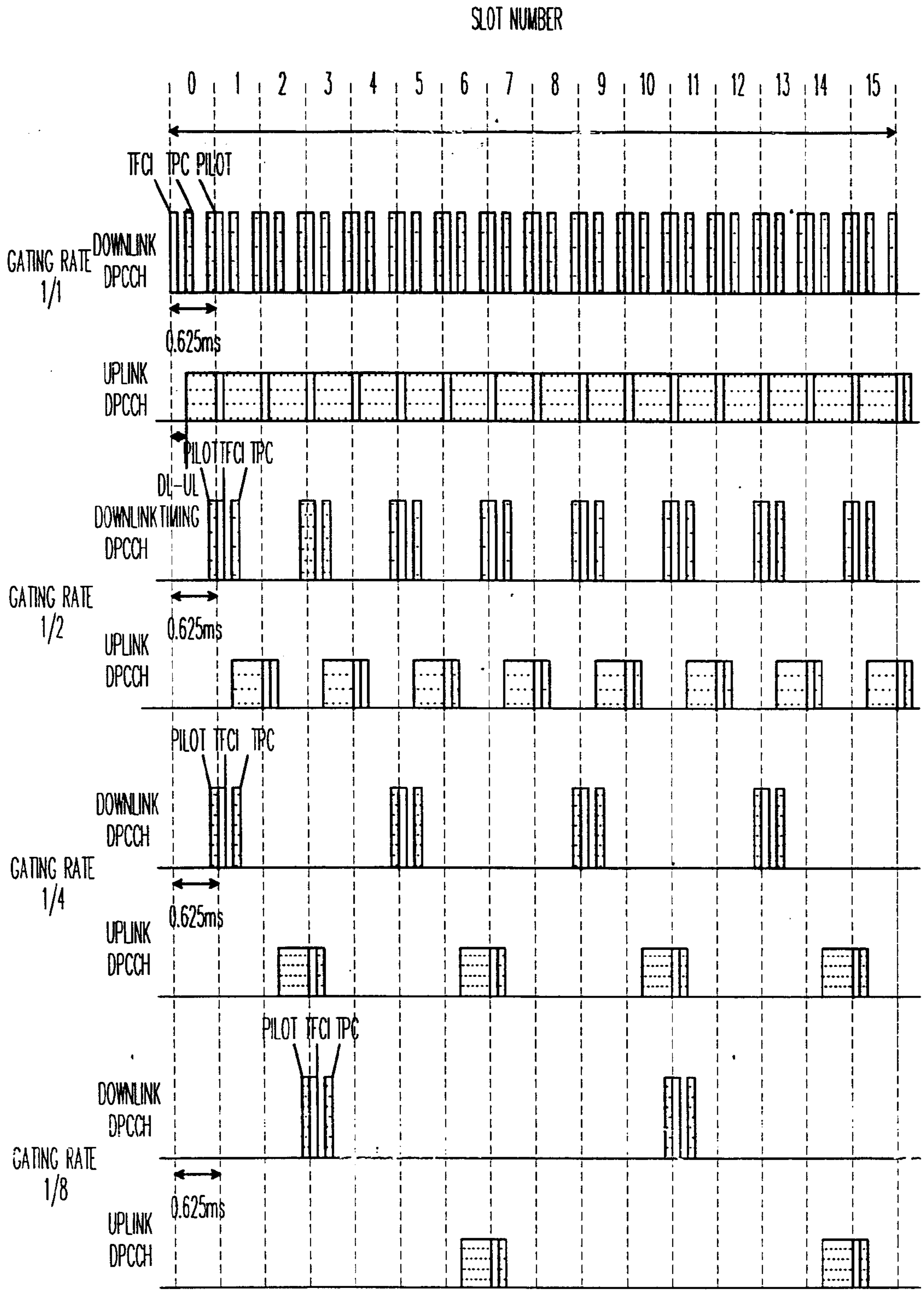

SLOT NUMBER
0
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
TFCI TPC PILOT
GATING RATE 1/1
DOWNLINK DPCCH
0.625ms
UPLINK DPCCH
PILOT TFCI TPC
DL-UL TIMING
DOWNLINK DPCCH
GATING RATE 1/2
0.625ms
UPLINK DPCCH
PILOT TFCI TPC
DOWNLINK DPCCH
GATING RATE 1/4
0.625ms
UPLINK DPCCH
PILOT TFCI TPC
DOWNLINK DPCCH
GATING RATE 1/8
0.625ms
UPLINK DPCCH

FIG. 11D

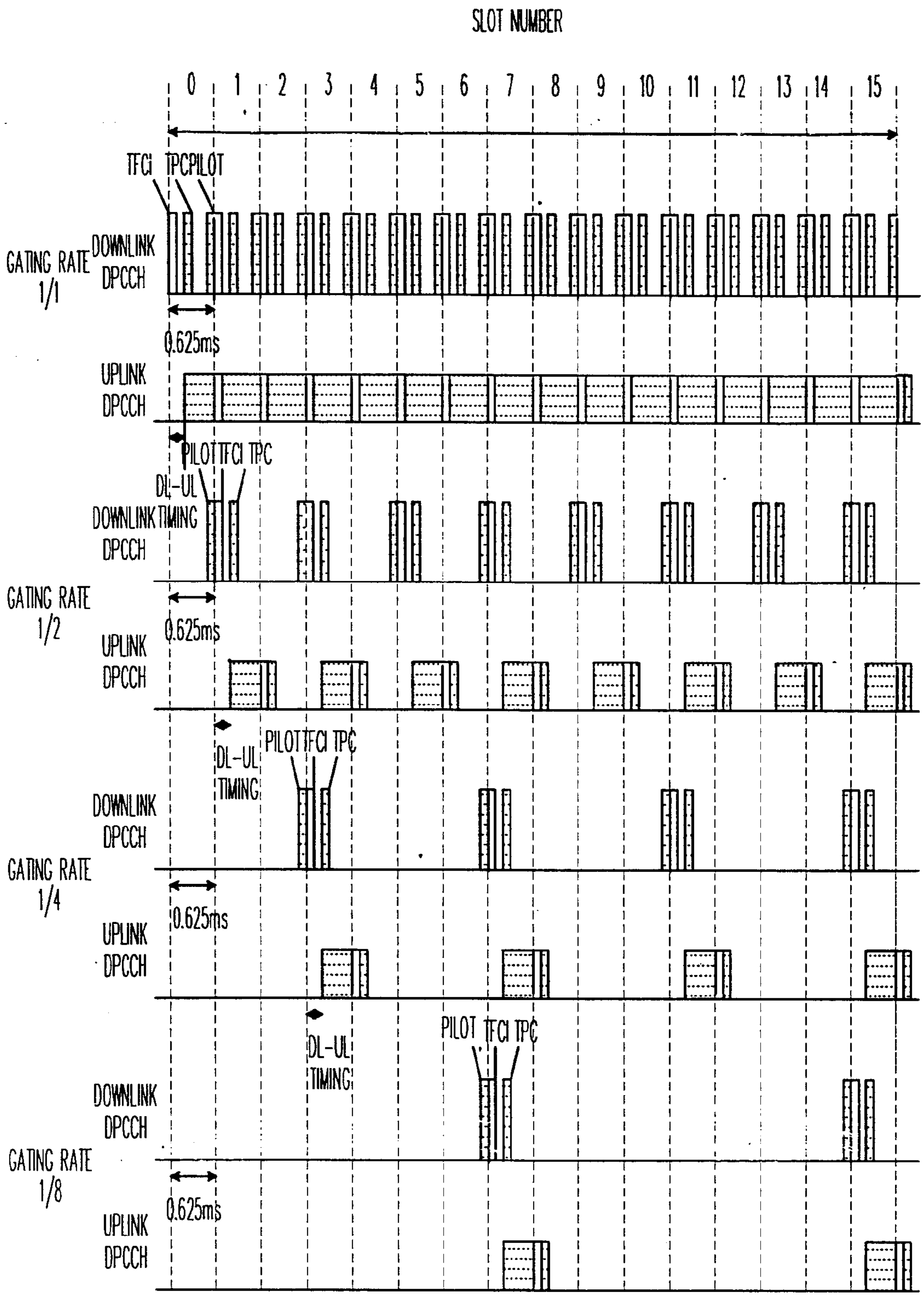

SLOT NUMBER
0
1
2
3
4
5
6
7
8
9
10
11
12
13
14
15
TFCI
TPC
PILOT
GATING RATE 1/1
DOWNLINK DPCCH
0.625ms
UPLINK DPCCH
DL-UL TIMING
GATING RATE 1/2
GATING RATE 1/4
GATING RATE 1/8

FIG. 11E

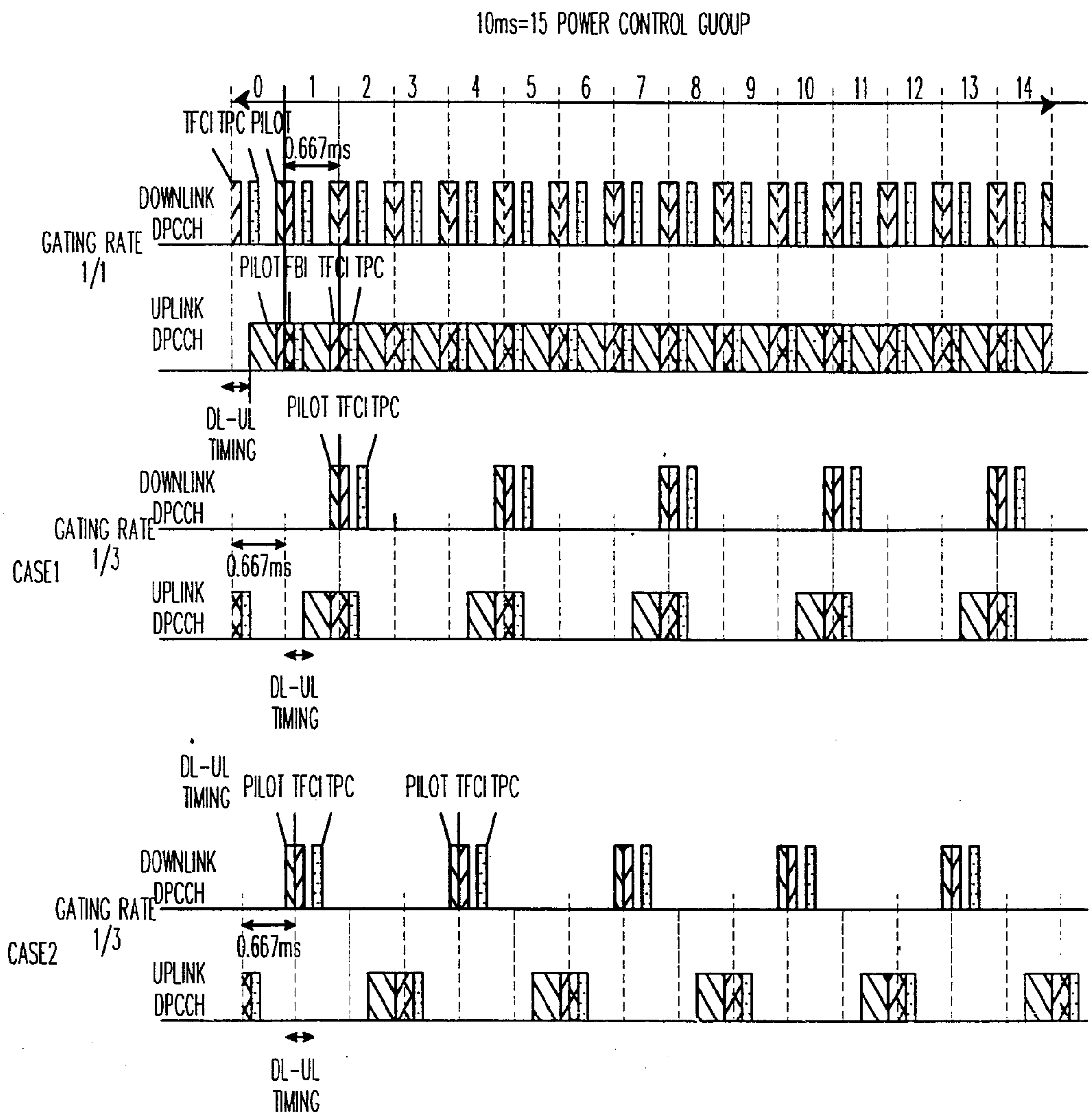

10ms=15 POWER CONTROL GUOUP
0
1
2
3
4
5
6
7
8
9
10
11
12
13
14
TFCI TPC PILOT
0.667ms
DOWNLINK DPCCH
GATING RATE 1/1
PILOT FBI TFCI TPC
UPLINK DPCCH
DL-UL TIMING
PILOT TFCI TPC
DOWNLINK DPCCH
GATING RATE 1/3
CASE1
0.667ms
UPLINK DPCCH
DL-UL TIMING
DL-UL TIMING
PILOT TFCI TPC
PILOT TFCI TPC
DOWNLINK DPCCH
GATING RATE 1/3
CASE2
0.667ms
UPLINK DPCCH
DL-UL TIMING

FIG. 12A

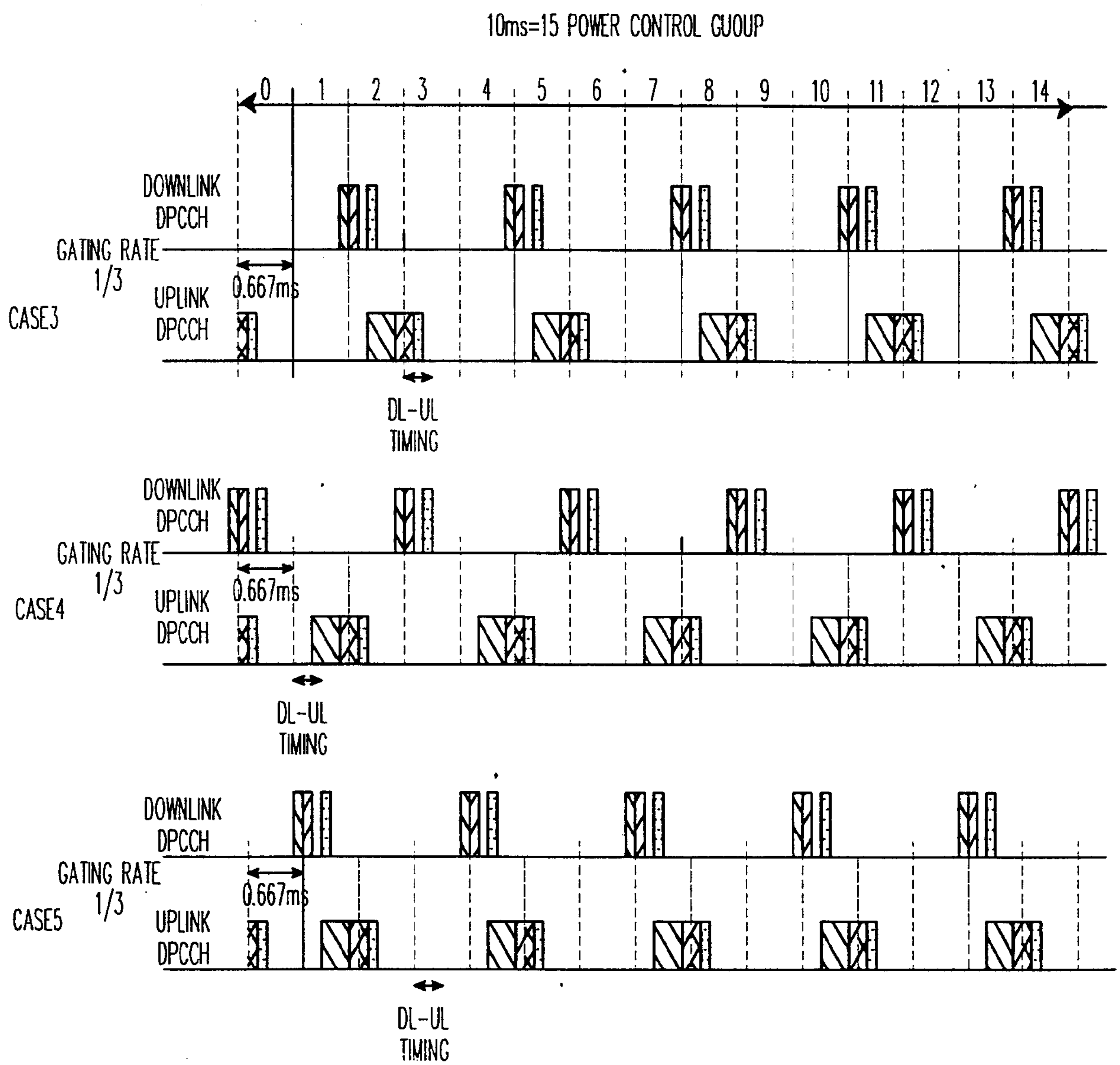

10ms=15 POWER CONTROL GUOUP
0
1
2
3
4
5
6
7
8
9
10
11
12
13
14
DOWNLINK
DPCCH
GATING RATE
1/3
0.667ms
UPLINK
DPCCH
CASE3
DL-UL
TIMING
DOWNLINK
DPCCH
GATING RATE
1/3
0.667ms
UPLINK
DPCCH
CASE4
DL-UL
TIMING
DOWNLINK
DPCCH
GATING RATE
1/3
0.667ms
CASE5
UPLINK
DPCCH
DL-UL
TIMING

FIG. 12A

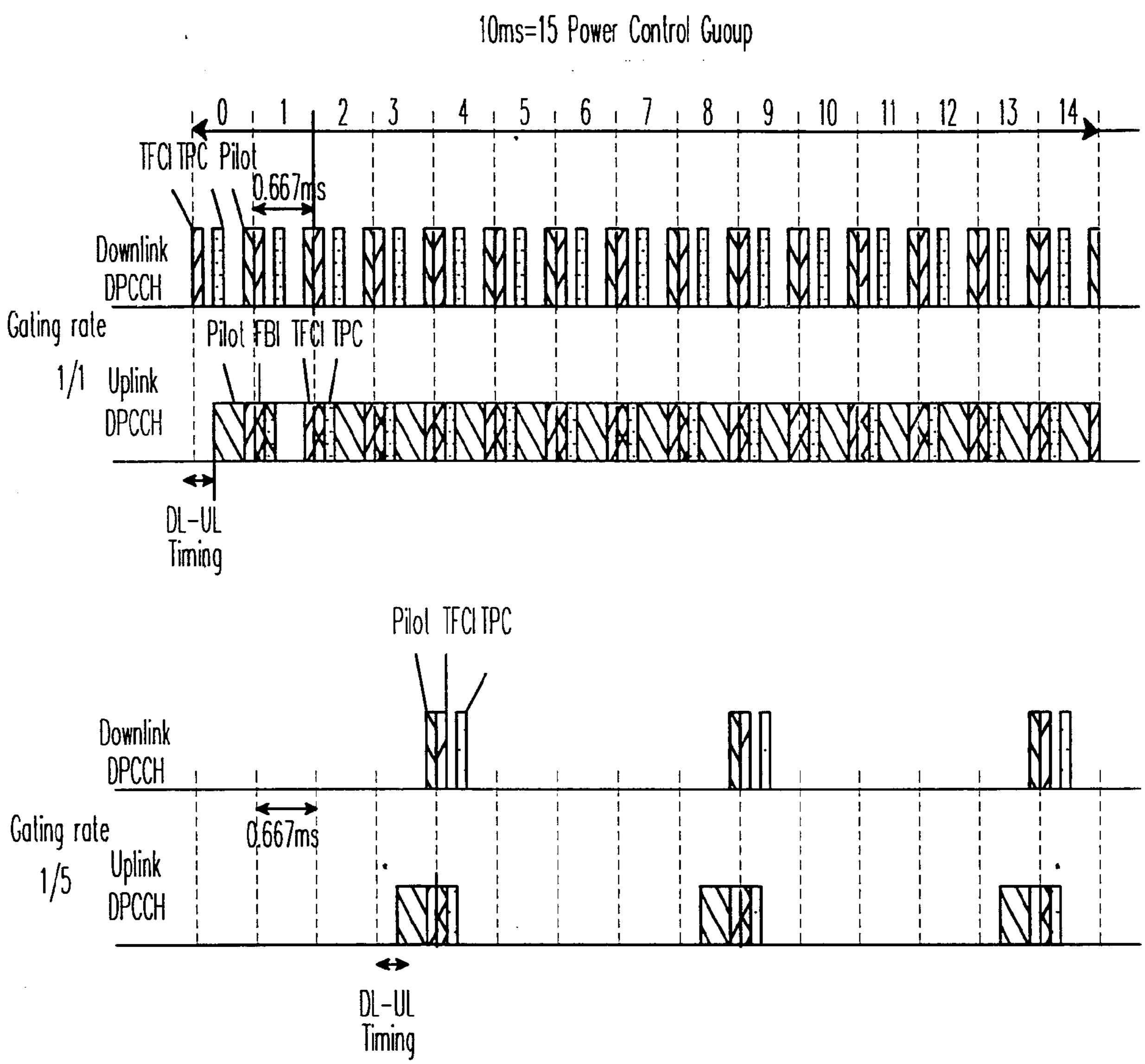

10ms=15 Power Control Guoup
0
1
2
3
4
5
6
7
8
9
10
11
12
13
14
TFCI TPC Pilot
0.667ms
Downlink DPCCH
Gating rate 1/1
Pilot FBI TFCI TPC
Uplink DPCCH
DL-UL Timing
Pilot TFCI TPC
Downlink DPCCH
Gating rate 1/5
0.667ms
Uplink DPCCH
DL-UL Timing

FIG. 12B

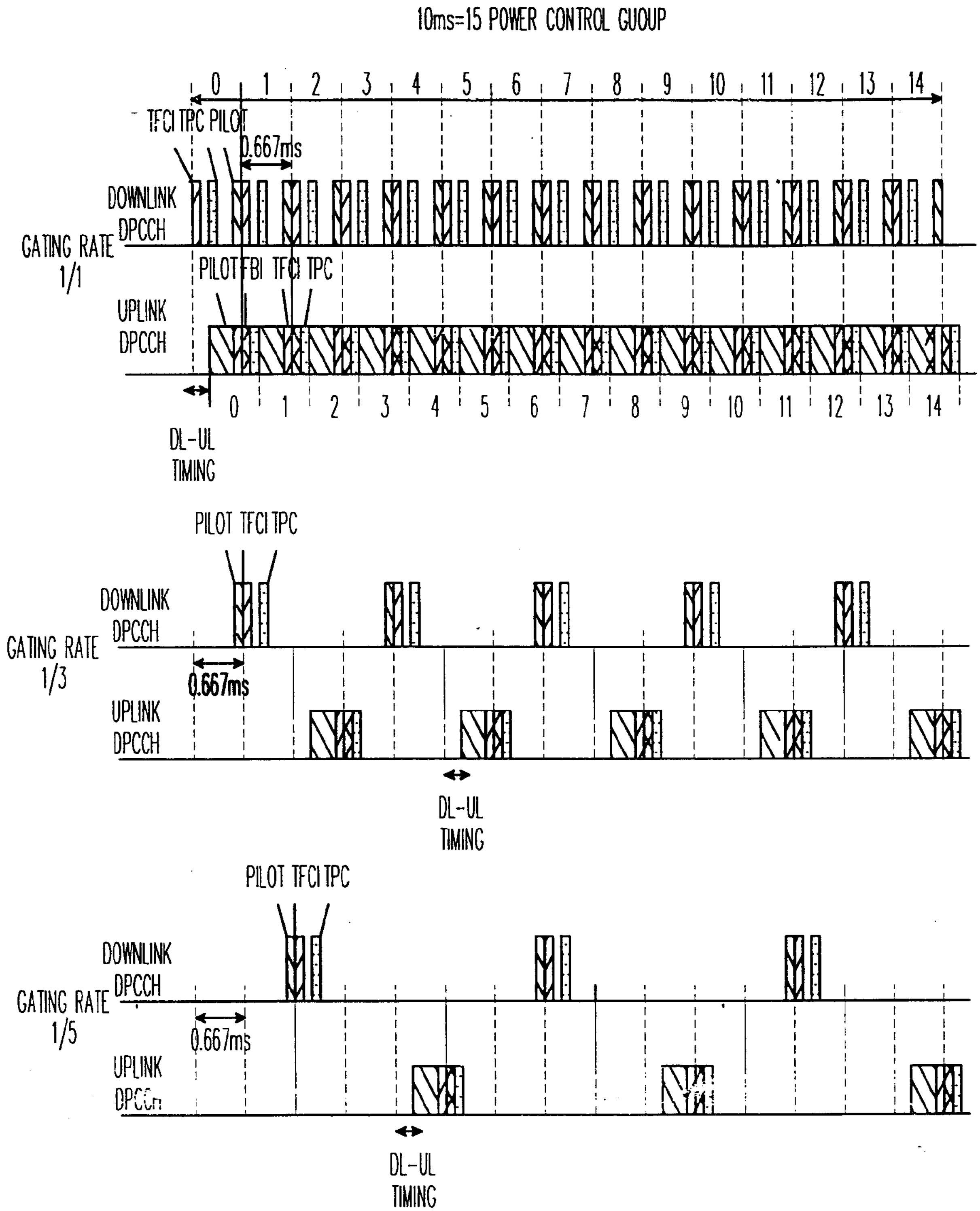

10ms=15 POWER CONTROL GUOUP
0
1
2
3
4
5
6
7
8
9
10
11
12
13
14
TFCI TPC PILOT
0.667ms
DOWNLINK DPCCH
GATING RATE 1/1
PILOT FBI TFCI TPC
UPLINK DPCCH
DL-UL TIMING
PILOT TFCI TPC
DOWNLINK DPCCH
GATING RATE 1/3
0.667ms
UPLINK DPCCH
DL-UL TIMING
PILOT TFCI TPC
DOWNLINK DPCCH
GATING RATE 1/5
0.667ms
UPLINK DPCCH
DL-UL TIMING

FIG. 12C

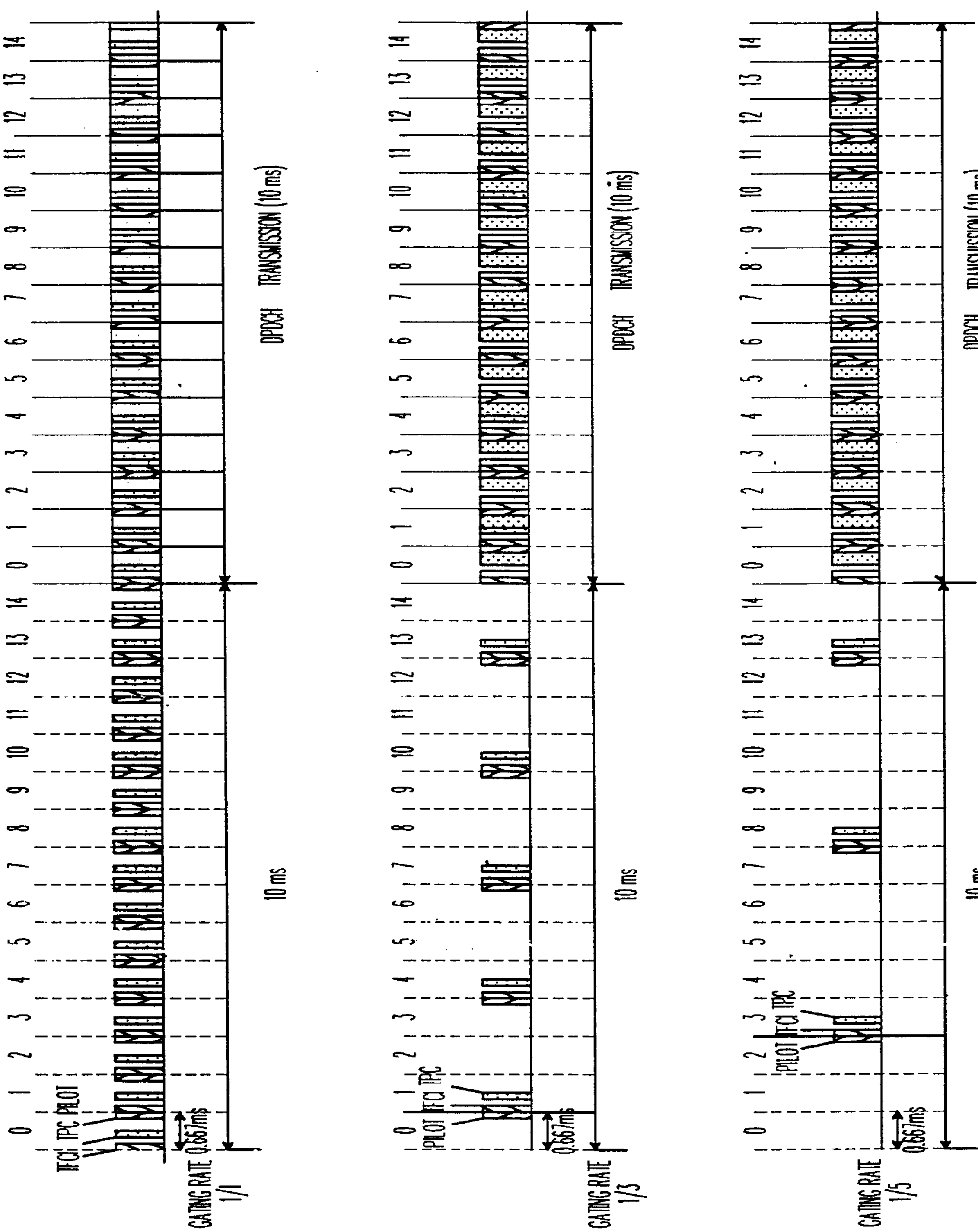
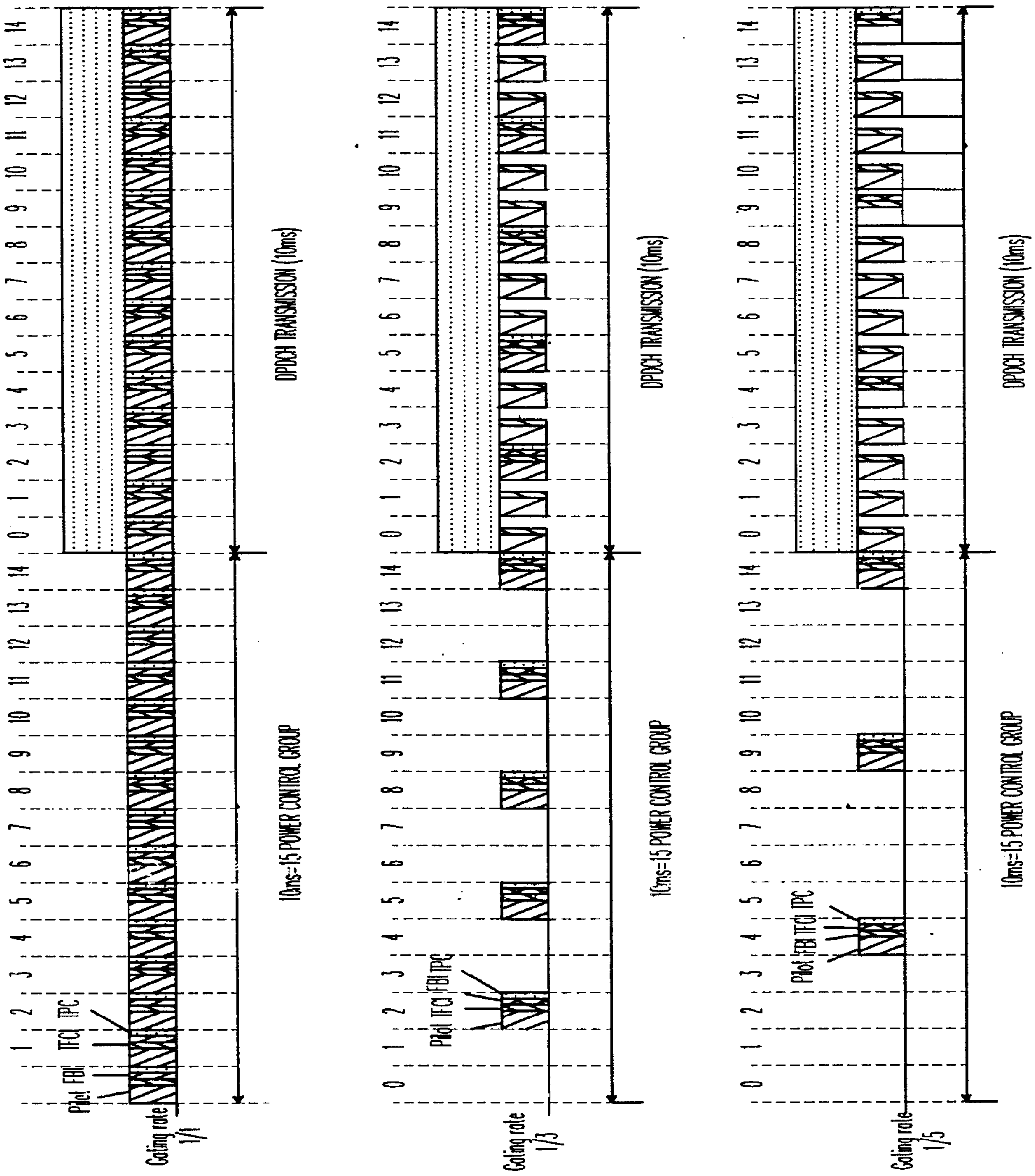

TFCI TPC PILOT
GATING RATE 1/1
0.667ms
10 ms
DPDCH TRANSMISSION (10 ms)
PILOT TFCI TPC
GATING RATE 1/3
GATING RATE 1/5

FIG. 12D

FIG. 12E


Pilot
FBI
TFCI
TPC
Gating rate
1/1
10ms=15 POWER CONTROL GROUP
DPDCH TRANSMISSION (10ms)
Pilot
TFCI
FBI
TPC
Gating rate
1/3
10ms=15 POWER CONTROL GROUP
DPDCH TRANSMISSION (10ms)
Pilot
FBI
TFCI
TPC
Gating rate
1/5
10ms=15 POWER CONTROL GROUP
DPDCH TRANSMISSION (10ms)

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
18 January 2001 (18.01.2001)

PCT

(10) International Publication Number
WO 01/05067 A1

(51) International Patent Classification[7]: H04B 7/26

(21) International Application Number: PCT/KR00/00740

(22) International Filing Date: 8 July 2000 (08.07.2000)

(25) Filing Language: English

(26) Publication Language: English

(30) Priority Data:
1999/28321 8 July 1999 (08.07.1999) KR

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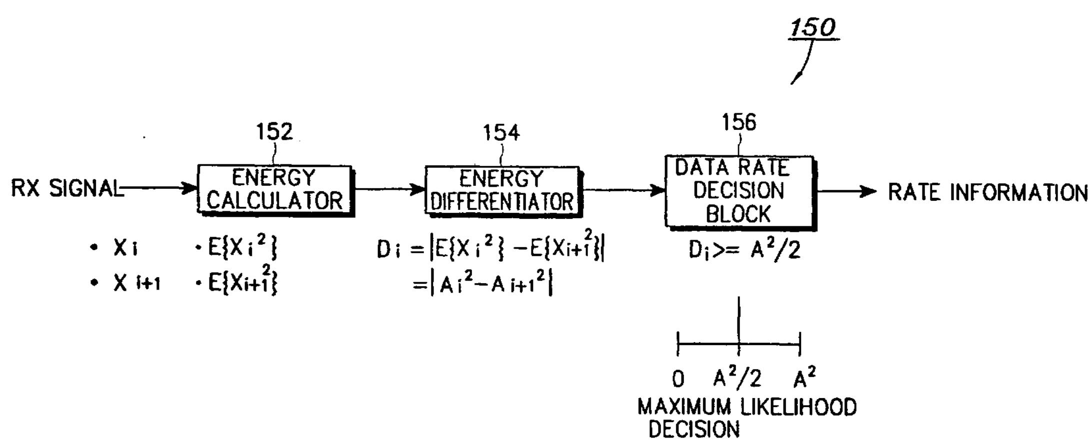
(81) Designated States *(national)*: AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, CA, CH, CN, CR, CU, CZ, DE, DK, DM, DZ, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, NO, NZ, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TR, TT, TZ, UA, UG, UZ, VN, YU, ZA, ZW.

(84) Designated States *(regional)*: European patent (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE).

Published:
— *With international search report.*

For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

(54) Title: DATA RATE DETECTION DEVICE AND METHOD FOR A MOBILE COMMUNICATION SYSTEM



(57) Abstract: A data rate detecting device detects a data rate for a received signal based on a variation of the energy for the respective received signals between the two adjacent intervals upon failure to receive information about the data rate, and performs channel decoding of the detected data rate information. First, the data rate detecting device divides an interval defined as between a lowest and highest one of a plurality of given data rates into m discriminating intervals. Then, the device calculates a difference between an average energy of received signals up to an i'th discriminating interval and an average energy of received signals for an (i+1)'th discriminating interval, wherein i is an integer is less than m. If the difference between the average energies is greater than or equal to a threshold value, the device determines that the received signal in the (i+1)'th discriminating interval is transmitted at a data rate corresponding to the i'th discriminating interval.

WO 01/05067 A1

DATA RATE DETECTION DEVICE AND METHOD FOR A MOBILE COMMUNICATION SYSTEM

BACKGROUND OF THE INVENTION

1. Field of the Invention

The present invention relates generally to a channel signal receiving device and method for a mobile communication system, and more particularly, to a device and method for detecting the data rate of a received signal.

2. Description of the Related Art

Code division multiple access (CDMA) mobile communication systems have developed from the conventional mobile communication standard, which focused on voice service, to the IMT-2000 standard, which provides high-speed data transmission. The IMT-2000 standard encompasses various services, including high quality voice, moving pictures, and Internet browsing. Communication links provided between a mobile station and a base station in the CDMA mobile communication system are generally classified into a downlink (DL), directing data to the mobile station from the base station, and an uplink (UL), directing data to the base station from the mobile station.

For voice or data transmission on the downlink or uplink, the data rate of the data may dynamically vary periodically, where the period is a predetermined time, e.g., 10 msec, which depends on the type of service. Usually, information about the data rate is transmitted to a receiver and used for decoding. However, in the event the receiver fails to receive the information about the data rate, the receiver has to detect rate of the received signal actually transmitted from the transmitter by analyzing the received signal. This procedure, where the receiver detects the data rate from the received signal, is called "blind rate detection (BRD)".

A description is provided herein below for a BRD operation according to the prior art which is performed in the case of voice transmission using convolutional codes for the purpose of forward error correction (FEC).

First, it is assumed that a set of data rates of voice data which a receiver (i.e., mobile station) uses to service a transmitter (i.e., base station) is designated as $R = \{R_1, R_2, \ldots, R_n\}$, where the data rates are listed in ascending order. To detect an actual data rate R_a reported by the transmitter, the receiver performs a Viterbi decoding of the data from the lowest data rate R_1 and then checks cyclic redundancy codes (CRC's). If the result of CRC check for R_1 is "good", there is a high probability that $R_a = R_1$, and R_a is determined as the actually transmitted data rate to be R_1. If the result of the CRC check for R_a is "bad", the receiver continues a Viterbi decoding of additional data up to the next data rate R_2, i.e., at a data rate $(R_2 - R_1)$, followed by CRC checks. As an attempt to reduce a false alarm potential of the BRD operation, the receiver checks an internal metric for Viterbi decoding, in addition to the CRC check.

As described above, the receiver first performs a Viterbi decoding and then a CRC check in order to detect a rate of convolution coded voice data. The BRD operation, however, is not easy to apply in the case of data transmission using turbo codes. This is because, unlike the Viterbi decoder, a turbo decoder has an internal turbo de-interleaver the type of which is dependent on the data rate. Specifically, when the result of CRC check at a given data rate is "bad", the turbo decoder has to repeat the data decoding process from the first data rate in order to check the CRC for a next data rate, while the Viterbi decoder has only to read additional data to the next data rate and then continue the data decoding. Another reason why the BRD operation is inadequate to the turbo decoder is in that the turbo decoding is usually performed iteratively, with the maximum number of iterations for a data rate being about 8 to 12, which leads to an increase in complexity of the decoder and which takes a long delay time when the iterative decoding is performed for CRC checks at all data rates.

SUMMARY OF THE INVENTION

It is, therefore, an object of the present invention to provide a device and method for detecting a data rate from a received signal upon failure to receive information about the data rate in a mobile communication system.

It is another object of the present invention to provide a device and method for detecting a data rate upon failure to receive information about the rate of turbo coded data.

It is yet another object of the present invention to provide a device and method for detecting a data rate not received during transmission of convolutional coded or turbo coded data.

It is still another object of the present invention to provide a device and method for reducing complexity of a data rate detecting operation upon failure to receive information about the data rate.

To achieve the above objects of the present invention, a data rate detecting device is provided that detects a data rate for a received signal based on a variation of the energy for the respective received signals between the two adjacent intervals upon failure to receive information about the data rate, and performs channel decoding of the detected data rate information.

The data rate detecting device first divides an interval defined as between a lowest and highest one of a plurality of given data rates into m discriminating intervals. Then, the device calculates a difference between an average energy of received signals up to an i'th discriminating interval and an average energy of received signals for an (i+1)'th discriminating interval, wherein i is an integer and is less than m. If the difference between the average energies is greater than or equal to a threshold, the device determines that the received signal in the (i+1)'th discriminating interval is transmitted at a data rate corresponding to the i'th discriminating interval.

BRIEF DESCRIPTION OF THE DRAWINGS

The above and other objects, features and advantages of the present invention will become more apparent from the following detailed description when taken in conjunction with the accompanying drawings in which:

Fig. 1 is a schematic block diagram illustrating a decoder for a mobile communication system including a data rate detector in accordance with the present invention;

Fig. 2 is a diagram illustrating a data rate detecting operation in accordance with the present invention;

Fig. 3 is a detailed block diagram illustrating the data rate detector shown in Fig. 1;

Fig. 4 is a flow chart illustrating the (i+1)'th interval data rate detecting operation in accordance with the present invention; and

Fig. 5 is a flow chart illustrating the i'th interval data rate detecting operation in accordance with the present invention.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENT

Turning to the drawings, a preferred embodiment of the present invention is described hereinbelow in detail with reference to the accompanying drawings. In the following description, well-known functions or constructions are not described in detail to avoid obscuring the invention in unnecessary detail.

Fig. 1 is a schematic block diagram of a decoder of a mobile station receiver in a mobile communication system including a data rate detector in accordance with the present invention. The invention is applicable to any CDMA mobile communication system, such as universal mobile telecommunication system (UMTS), CDMA2000, etc.

Referring to Fig. 1, a de-interleaver 110 de-interleaves a received signal to generate a de-interleaved signal (symbol) X_k. Discontinuous transmission (DTX) bit extractor 120 extracts, from the de-interleaved signal X_k, bits indicating a discontinuous transmission mode received from a base station in a discontinuous transmission mode of the mobile communication system. Data rate detector 150 detects a variable data rate of the received signal (symbol) X_k de-interleaved at the de-interleaver 110, ultimately detecting the rate of the received data upon the failure to receive information regarding the data rate. Specifically, the data rate detector 150 measures variations of energy for each received signal in two adjacent intervals and detects the data rate of the received signal based on the result of detection. The information about the data rate detected at the data rate detector 150 is applied to a rate matching block 130 and a channel decoder 140. The rate matching block 130 receives the de-interleaved symbols to perform a reverse process of puncturing, i.e. symbol insertion, and a reverse process of repetition, i.e. symbol combining, thus generating rate-matched symbols. Channel decoder 140 decodes the rate-matched symbols received from the rate matching block 130. The channel decoder 140 may be implemented with a convolutional decoder or a turbo decoder. The rate matching block 130 and the channel decoder 140 use the data

rate information received from the data rate detector 150 to perform rate matching and channel decoding operations.

Fig. 2 is an illustration for explaining a data rate detecting operation of the present invention performed at the data rate detector 150 shown in Fig. 1.

First, it is assumed that the number of symbols received at the mobile station receiver varies in the order of R_1, R_2, R_3, R_4 and R_5 over time, as shown in Fig. 2. A change in the number of symbols by the interval, e.g., 10 msec means that the data rate is variable. Thus it should be noted that the term "the number of symbols" is substantially interchangeable with the term "data rate".

Fig. 2 shows a case where the base station transmitter correctly transmits data for intervals 1 to 4 but fails to transmit data between intervals 4 and 5. The data symbols in the transmission intervals 1 to 4 are de-interleaved at the de-interleaver 110 shown in Fig. 1 and stored in an internal buffer of the DTX(Discontinuous Transmission) bit extractor 120. Between the intervals 4 and 5, the base station transmitter sends DTX bits in a DTX mode. For such a DTX interval, the base station transmitter disables the transmission power and only an additive white Gaussian noise (AWGN) exists. So, the data rate is R_4 for the DTX interval 5. As such, the present invention uses a fundamental principle that involves determination of a presence of the data in substantially non-transmission intervals for data or data rate information, and ultimately detection of the data rate.

Now, a detailed description will be given to the principle of the data rate detection according to the present invention.

Expediently, it is assumed that there are two data rates R_1 and R_2. In such a case, the following equations may be used in order to determine, without receiving any data rate information, whether a signal has been transmitted at R_1 or R_2. When the received signal from bit position 1 to bit position R_1 is X_1, and the received signal from bit position (R_1+1) to bit position R_2 is X_2, the signals X_1 and X_2 are expressed by:

[Equation 1]

$$X_1 = A_1 \times a_1 + n_1$$

$$X_2 = A_2 \times a_2 + n_2$$

In Equation 1, A_1 and A_2 represent transmission power levels of the signals transferred from the base station transmitter to the mobile station receiver and correspond to $\pm A$ in the presence of the signals or "0" for DTX; a_1 and a_2 represent Rayleigh random variables having a probability function of $p(a_1) = 2 \times a_1 \times \exp(-a_1^2)$ or $p(a_2) = 2 \times a_2 \times \exp(-a_2^2)$, respectively; and n_1 and n_2 represent AWGN random variables with mean "0" and variance σ^2. If the noise variance of the transmission channel is σ^2, the interval-based energy (power) of the received signal is given by:

[Equation 2]

$$E\{X_1^2\} = A_1^2 + \sigma^2$$

$$E\{X_2^2\} = A_2^2 + \sigma^2$$

The differentiation equation of the energies $E\{X_1^2\}$ and $E\{X_2^2\}$ of the received signals gives D_1 as expressed by:

[Equation 3]

$$D_1 = \left|E\{X_1^2\} - E\{X_2^2\}\right| = \left|A_1^2 - A_2^2\right|$$

In Equation 3, if $A_1^2 = A_2^2$, D_1 is "0"; otherwise, if $A_2^2 = 0$ (i.e., for DTX), D_1 is "A_1^2". Namely, when the actual data rate is R_2, D_1 nearly reaches "0"; otherwise, when the actual data rate is R_1, D_1 becomes almost "A_1^2".

The above equations can be applied only if the secondary probability characteristic, average deviation σ^2 is known irrespective of the probability functions $p(a_1)$ and $p(a_2)$ of the Rayleigh random variables. It is of cause assumed that the random variables is not time varying. For reference, the differentiation result of the energies of the received signals $D_1 = \left|E\{X_1^2\} - E\{X_2^2\}\right|$ can be calculated from a given interval-based energy of the individual received signals. The most important variable in determining D_1 may be the accumulation of data sufficient to determine the average energy value. An accurate data rate may be determined when the minimum data rate R_1 is 32 kbps, i.e., the data transmitted in the 10msec frame interval is more than 320 bits.

The above-stated data rate detecting operation can be generalized as follows.

First, it is assumed that a set of serviceable data rates is designated as $R = \{R_1, R_2, \ldots, R_n\}$, in which the data rates are listed in the ascending order. Information about the serviceable data rates is called "transport format set (TFS)" given to the mobile station by the base station in a call setup phase. If information about n data rates is given, one interval is first assigned to the largest data rate R_n and (n-1) intervals are assigned to the other data rates. To be differentiated from the interval assigned to the largest data rate R_n, (n-1) intervals for the other data rates are defined as discriminating intervals. The data rate of the received signal for the individual is detectable. For instance, an average energy of the received signals up to the i'th discriminating interval is subtracted from an average energy of the received signals up to the (i+1)'th discriminating interval. The resulting subtracted value is compared to a predetermined threshold to detect the data rate of the received signal for the (i+1)'th interval.

Now, the operation of detecting the data rate of the received signal for the (i+1)'th interval is described in connection with generalized expressions as follows. A received signal up to the i'th interval designated as X_i can be defined as:

[Equation 4]

$$X_i = A_i \times a_i + n_i$$

In Equation 4, A_i represents the transmission power level of the base station transmitter and correspond to $\pm A$ in the presence of the signal or "0" for DTX; and a_i and n_i represent the Rayleigh random variable and the AWGN random variable as defined above, respectively. From Equation 3, a decision criterion can be defined as in Equation 5 below, from 1 to n. When the received signal up to the i'th interval is X_i and a received signal up to the (i+1)'th interval is X_{i+1}, the differentiation result of the energies $E\{X_i^2\}$ and $E\{X_{i+1}^2\}$ of the received signals gives D_i as expressed by:

[Equation 5]

$$D_i = \left|E\{X_i^2\} - E\{X_{i+1}^2\}\right| = \left|A_i^2 - A_{i+1}^2\right|$$

In Equation 5, if the data are continuously transmitted up to the (i+1)'th interval, i.e., $A_i^2 = A_{i+1}^2$, then D_i is "0"; otherwise, if the data are transmitted up to the i'th interval but not transmitted from the i'th to the (i+1)'th interval (for DTX), i.e., $A_{i+1}^2 = 0$, then D_i is "A_i^2". Therefore, during DTX ($A_{i+1}^2 = 0$), the mobile

station receiver searches for the initial index i and considers the corresponding R_i as the actual data rate for the received data from the base station transmitter.

Fig. 3 is a schematic block diagram of the data rate detector 150 shown in Fig. 1, in which the data rate detector 150 comprises an energy calculator 152, an energy differentiator 154 and a data rate decision block 156.

Referring to Fig. 3, the energy calculator 152 measures energy E_i for a received signal X_i up to the i'th interval and energy E_{i+1} for a received signal X_{i+1} from the i'th interval to the (i+1)'th interval. Namely, the energy calculator 152 accumulates the received signals up to the i'th interval and the received signals up to the (i+1)'th interval to calculate energies E_i and E_{i+1} for the respective received signals X_i and X_{i+1} according to Equation 6 below, which is used to calculate energy E_{i+1} for the received signal X_{i+1}.

[Equation 6]

$$E_{i+1} = \frac{1}{R_{i+1} - R_i} \sum_{k=R_i}^{R_{i+1}} X_k^2 dk$$

The energy differentiator 154 calculates a difference (D_i) between energy $E\{X_i^2\}$ in the i'th interval and energy $E\{X_{i+1}^2\}$ in the (i+1)'th interval, as obtained in Equation 6. The difference between the energies $E\{X_i^2\}$ and $E\{X_{i+1}^2\}$ may be expressed as a difference between the squares of the transmission power levels, as defined in Equations 3 and 5, i.e., a difference between a square A_i^2 of the transmission power level of a received signal for the i'th interval in the i'th interval, and a square A_{i+1}^2 of the transmission power level of a received signal for the (i+1)'th interval. The data rate decision block 156 determines the rate of the transmission data using the energy difference D_i calculated at the energy differentiator 154. If D_i is a desired value A_i^2 as in Equation 5, the data rate decision block 156 determines the data rate R_i for the i'th interval as the rate of the presently transmitted data.

However, considering the actual channel environment, it is impossible that the energy difference between the two intervals as designated by D_i is "0" or A_i^2. That is, the difference D_i itself is a probability variable, where the conditional expectation of D_i satisfies $E\{D_i | A_i^2 = A_{i+1}^2\} = 0$ and $E\{D_i | A_i^2 \neq A_{i+1}^2\} = A^2$. Thus, the data rate decision block 156 compares the energy difference D_i between the two adjacent intervals with a threshold value to determine the data rate. More particularly, the data rate decision block156 determines the data rate R_i for the

previous interval, the i'th interval as the data rate for the current interval when the energy difference D_i between the two adjacent intervals is less than or equal to the threshold value. The threshold value can be designated as a medium value between "0" and A^2, i.e., $A^2/2$ according to a maximum likelihood (ML) principle. Here, A denotes the transmission power level of the received signal from the base station transmitter and $A^2/2$ is half the transmission power level of the received signal. The information about the data rate determined by the data rate decision block 156 is applied to the rate matching block 130 and the channel decoder 140, as shown in Fig.1.

The flow chart of Figs. 4 and 5 illustrate a data rate detecting operation using the above equations which is performed at the data rate detector 150 shown in Fig. 3. Fig. 4 is a flow chart illustrating an operation of detecting the data rate for the (i+1)'th interval from the energies of the received signals for the two adjacent intervals, the i'th and (i+1)'th intervals. Fig. 5 is a flow chart illustrating a general operation of detecting the data rate for the i'th interval.

Referring to Fig. 4, the data rate detector 150 shown in Fig. 1 calculates the energy difference D_i between the two adjacent intervals for each iteration and compares the energy difference D_i with a threshold value $A^2/2$. The data rate detector150 estimates the data rate R_i for the i'th interval as an actual data rate R_{est}, in step 405, when the energy difference D_i is greater than or equal to the threshold value.

Specifically, the energy calculator 152 shown in Fig. 3 accumulates received signal X_i between the (i-1)'th interval and the i'th interval in step 401, and calculates energy $E\{X_i^2\}$ for the received signal X_i in step 402. Also, the energy calculator 152 accumulates received signal X_{i+1} between the i'th interval and the (i+1)'th interval and calculates energy $E\{X_{i+1}^2\}$ for the received signal X_{i+1} in step 402. The energy differentiator 154 calculates an energy difference between the two adjacent intervals, in step 403. That is, the energy differentiator 154 determines the energy difference between the two intervals as $D_i = \left|E\{X_i^2\} - E\{X_{i+1}^2\}\right|$. As previously stated, the energy difference can also be expressed as $D_i = \left|A_i^2 - A_{i+1}^2\right|$. In step 404, the data rate decision block 156 compares the energy difference between the two adjacent intervals with a threshold value, i.e., it determines whether the energy difference D_i is greater than or equal to the threshold value $A^2/2$. When the energy difference D_i is greater than or equal to the threshold value $A^2/2$, the data rate decision block 156 estimates the data rate R_i for the i'th interval

as the actual data rate R_{est} for the current (i+1)'th interval, in step 405. The estimated data rate is provided to the DTX bit extractor 120, the rate matching block 130 and the channel decoder 140, as shown in Fig. 1, and used for rate matching and channel decoding operations.

Referring to Fig. 5, in step 501, the data rate detector initializes the search interval i to "1" and the average power (energy) for the previous interval $E\{X_{i-1}^2\}$ to "0". The energy calculator 152 shown in Fig. 3 calculates, in step 502, the average power for the search interval 1, i.e., first calculates the average power for the current interval $E\{X_i^2\}$. In step 503, the energy differentiator 154 calculates (a second calculation) an energy difference between the previous interval and the current interval according to discriminating equation D_{i-1}. If the data rate decision block 156 determines in step 504 that the result of discriminating equation D_{i-1} is greater than or equal to the threshold value $A^2/2$ (where, the data rate means "0" kbps as i = 1), the data rate decision block 156 estimates the data rate for the current interval R_{est} as the data rate for the previous interval (R_{i-1}) in step 508.

Otherwise, i.e., if it is determined in step 504 that the result of discriminating equation D is less than the threshold value $A^2/2$, the data rate decision block 156 stores the average power $E\{X_i^2\}$ for the current interval in the average power $E\{X_{i-1}^2\}$ for the previous interval in step 505, and increases i by one for searching the next interval in step 506. The energy calculator 162 in step 507 calculates (a third calculation) the average power in the interval i+1 and then stores the calculated average power in the average power $E\{X_i^2\}$ for the current interval. The process returns to step 503 to calculate the discriminating equation D_{i-1} based on the average power $E\{X_i^2\}$ and compares in step 504 the result value of the discriminating equation D_{i-1} with the threshold value.

While repeating the above procedures, when it is determined as $D \geq A^2/2$ in step 504, the data rate decision block 156 estimates the data rate R_{est} of the current interval as the data rate R_{i-1} up to the previous interval.

As described above, the present invention estimates a data rate for a received signal prior to a decoding operation even when no information about the data rate is received from the base station transmitter, which reduces the complexity as compared to the conventional BRD operation which detects the data rate after Viterbi decoding and the CRC check. The present invention thereby reduces the complexity in detecting the rate of turbo-encoded data without a need of a rate-based decoding operation, in the worse case, as often as the maximum number of iterations.

Furthermore, the present invention determines the data rate using only accumulated statistics, irrespective of the type of the channel encoder, and is thus compatible with any channel encoder. For example, even with a convolutional encoder is used, the present invention makes it possible to estimate the data rate with reliability for a frame whose data rate is not less than a threshold value.

While the invention has been shown and described with reference to a certain preferred embodiment thereof, it will be understood by those skilled in the art that various changes in form and details may be made therein without departing from the spirit and scope of the invention as defined by the appended claims.

WHAT IS CLAIMED IS:

1. A method for detecting a data rate in a mobile communication system, comprising the steps of:

dividing an interval defined as between a lowest and highest one of a plurality of predetermined data rates into m discriminating intervals, wherein m is an integer; and

calculating a difference between an average energy of received signals up to an i'th discriminating interval and an average energy of received signals for an (i+1)'th discriminating interval, wherein i is an integer and is less than m; and

when the difference between the average energies is greater than or equal to a threshold value, determining that the received signal in the (i+1)'th discriminating interval is transmitted at a data rate corresponding to the i'th discriminating interval.

2. The method as claimed in claim 1, wherein the threshold value is defined as $A^2/2$, wherein A represents a transmission power level of the received signal up to the i'th discriminating interval.

3. A device for detecting a data rate in a mobile communication system, in which an interval is defined as between a lowest and highest one of a plurality of given data rates being divided into m discriminating intervals, wherein m is an integer, the device comprising:

an energy calculator for calculating an average energy of received signals up to an i'th discriminating interval and an average energy of received signals for an (i+1)'th discriminating interval, wherein i is an integer and is less than m;

an energy differentiator for calculating a difference between the average energy of received signals up to the i'th discriminating interval and the average energy of received signals for the (i+1)'th discriminating interval; and

a data rate decision block for determining a data rate corresponding to the i'th discriminating interval as a data rate for the received signal in the (i+1)'th discriminating interval, when the difference between the average energies calculated in the energy calculator is greater than a threshold value.

4. The device as claimed in claim 3, wherein the threshold value is defined as $A^2/2$, wherein A represents a transmission power level of the received signal up to the i'th discriminating interval.

5. A method for detecting a data rate in a mobile communication system, in which a base station has previously provided a mobile station with information about a plurality of data rates variably serviceable and the mobile station detects one of the plurality of data rates as a data rate for a received signal, the method comprising the steps of:

(a) dividing an interval defined as between a lowest and highest one of the plurality of data rates into m discriminating intervals, wherein m is an integer; and

(b) calculating an average energy of a received signal corresponding to a first discriminating interval out of the m discriminating intervals;

(c) calculating an average energy of a received signal corresponding to a second discriminating interval next to the first discriminating interval;

(d) calculating a difference between the average energies obtained in steps (b) and (c); and

(e) estimating that the received signal for the second discriminating interval is transmitted at a data rate corresponding to the received signal for the first discriminating interval, when the difference between the average energies is greater than or equal to a threshold value, or

setting the first discriminating interval as the next discriminating interval when the difference between the average energies is greater than or equal to the threshold value,

the steps (b) to (e) for the received signals up to the newly set discriminating interval being repeated until the difference exceeds the threshold value.

FIG. 1

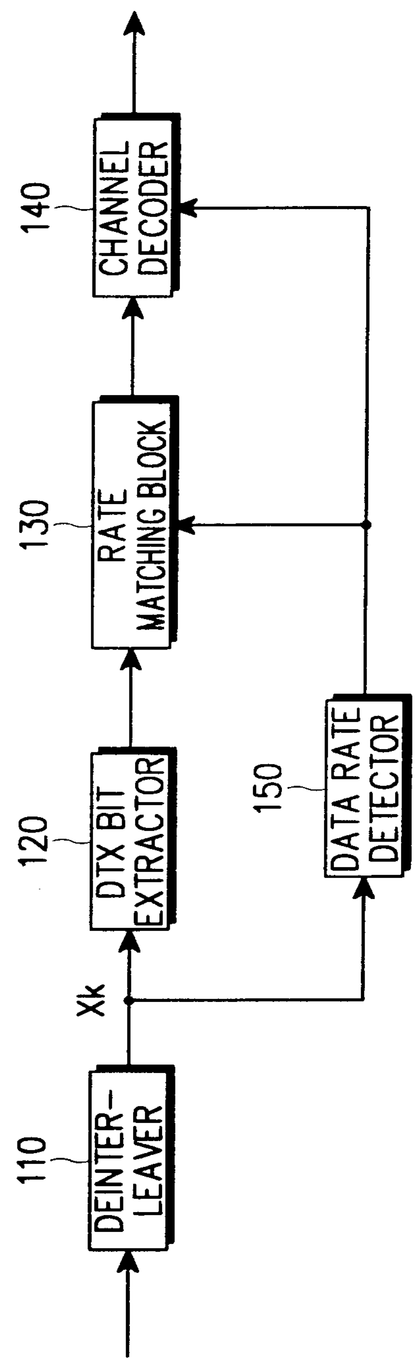

110
DEINTER-LEAVER
Xk
120
DTX BIT EXTRACTOR
150
DATA RATE DETECTOR
130
RATE MATCHING BLOCK
140
CHANNEL DECODER

FIG. 2

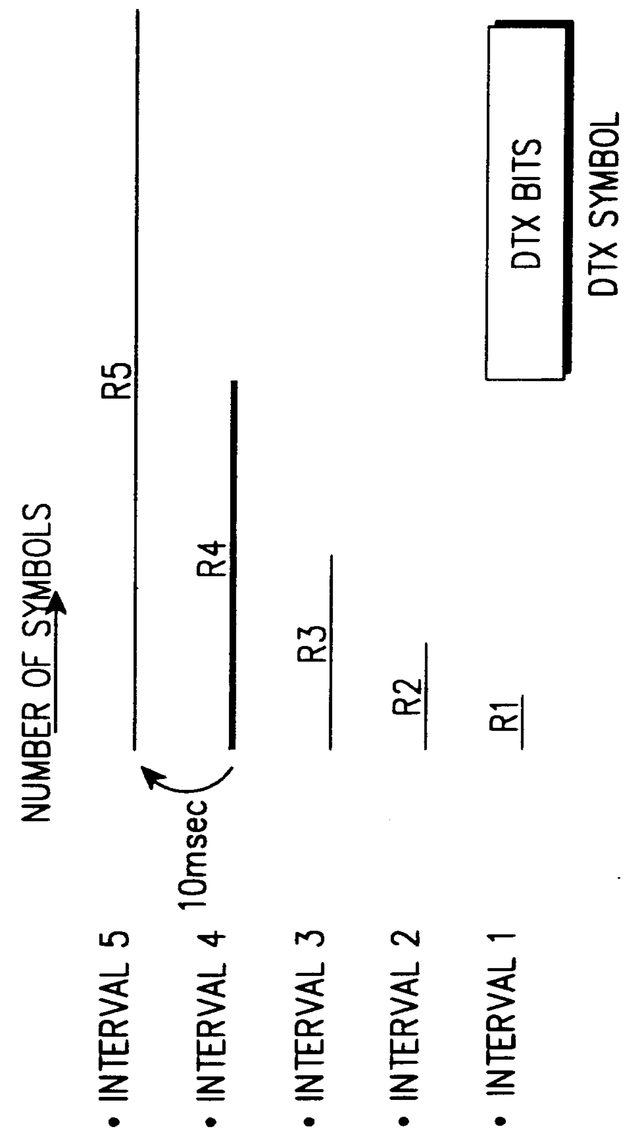
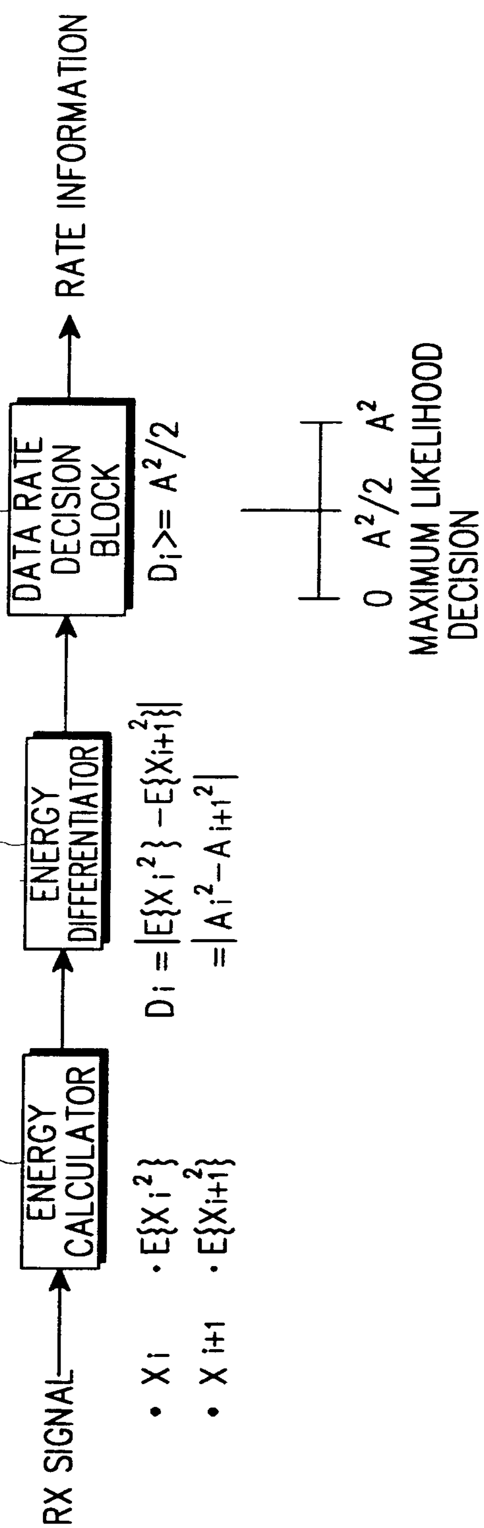

• INTERVAL 5
• INTERVAL 4
• INTERVAL 3
• INTERVAL 2
• INTERVAL 1
10msec
NUMBER OF SYMBOLS
R1
R2
R3
R4
R5
DTX BITS
DTX SYMBOL

FIG. 3


150
RX SIGNAL
• X_i
• X_{i+1}
152
ENERGY CALCULATOR
· $E\{X_i^2\}$
· $E\{X_{i+1}^2\}$
154
ENERGY DIFFERENTIATOR
$D_i = |E\{X_i^2\} - E\{X_{i+1}^2\}|$
$= |A_i^2 - A_{i+1}^2|$
156
DATA RATE DECISION BLOCK
$D_i >= A^2/2$
RATE INFORMATION
0
$A^2/2$
A^2
MAXIMUM LIKELIHOOD DECISION

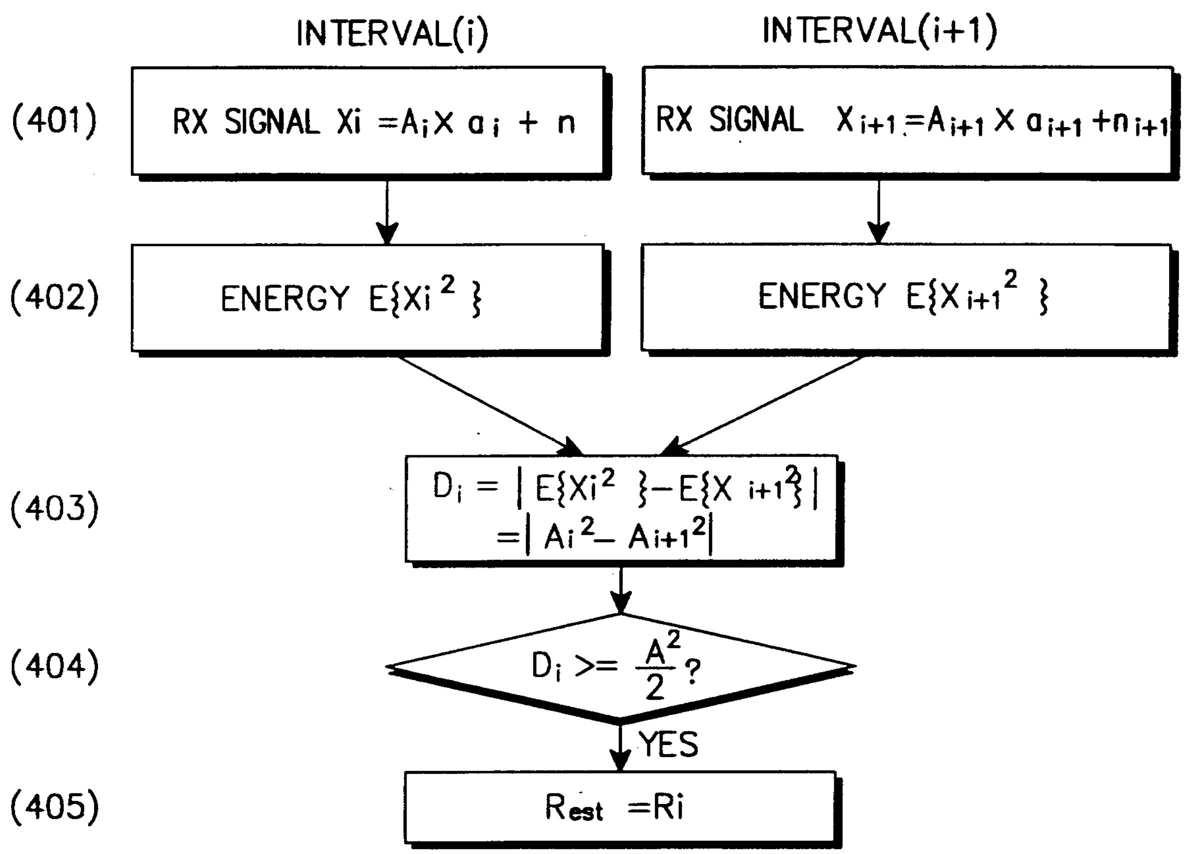

INTERVAL(i)
INTERVAL(i+1)
(401)
RX SIGNAL Xi =AiX ai + n
RX SIGNAL Xi+1=Ai+1 X ai+1 +ni+1
(402)
ENERGY E{Xi 2 }
ENERGY E{Xi+1 2 }
(403)
Di = | E{Xi2 }−E{X i+12}|
=| Ai2− Ai+12|
(404)
Di >= A2/2 ?
YES
(405)
Rest =Ri

FIG. 4

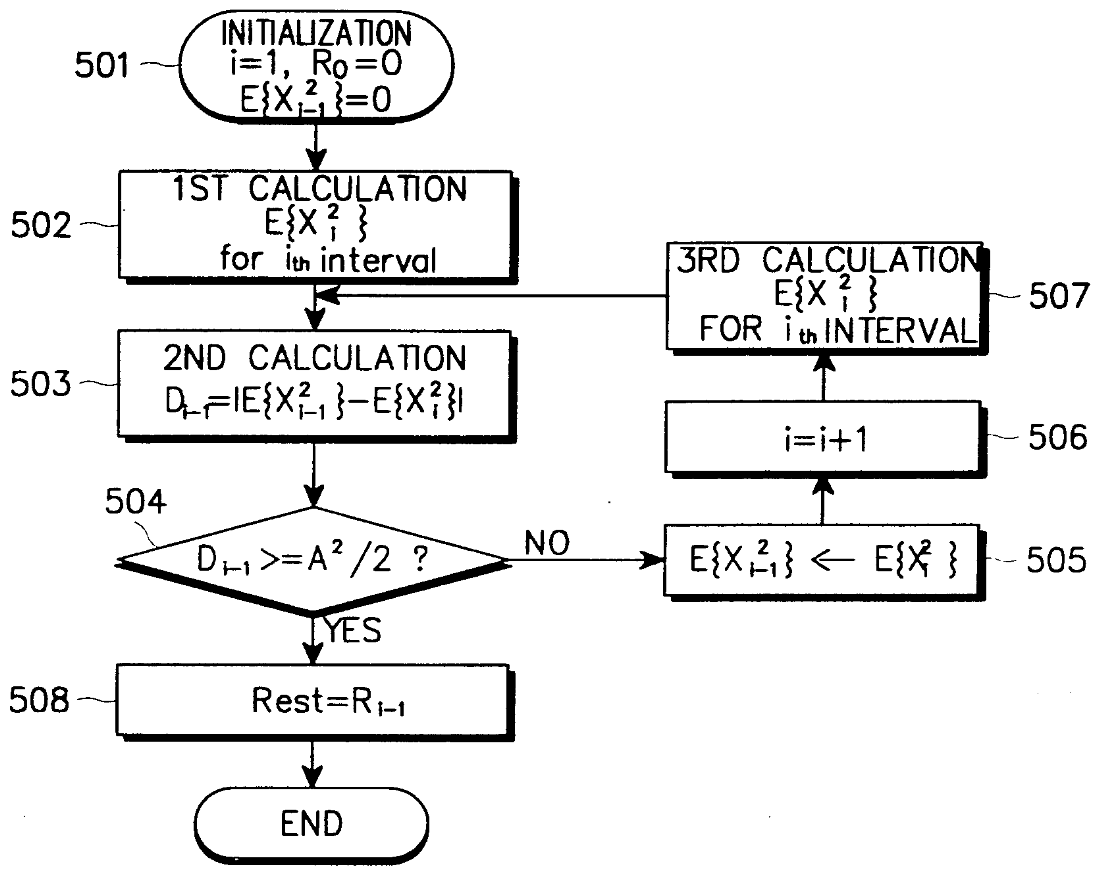

INITIALIZATION
i=1, R_0=0
$E\{X_{i-1}^2\}=0$
501
1ST CALCULATION
$E\{X_i^2\}$
for i_{th} interval
502
2ND CALCULATION
$D_{i-1}=|E\{X_{i-1}^2\}-E\{X_i^2\}|$
503
504
$D_{i-1}>=A^2/2$?
NO
YES
$E\{X_{i-1}^2\} \leftarrow E\{X_i^2\}$
505
i=i+1
506
3RD CALCULATION
$E\{X_i^2\}$
FOR i_{th} INTERVAL
507
Rest=R_{i-1}
508
END

FIG. 5

INTERNATIONAL SEARCH REPORT

International application No.
PCT/KR00/00740

A. CLASSIFICATION OF SUBJECT MATTER

IPC7 H04B 7/26

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimun documentation searched (classification system followed by classification symbols)

IPC7 H04B, H04L

Documentation searched other than minimun documentation to the extent that such documents are included in the fileds searched

Korean Patents and applications for inventions since 1975
Korean Utility models and applications for Utility models since 1975

Electronic data base consulted during the intertnational search (name of data base and, where practicable, search trerms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US 5566206 A (Qualcomm INC.) 15 October 1996 See the whole document	1,3,5
A	US 5671255 A (Motorola INC.) 23 September 1997 See the whole document	1,3,5
A	US 5751725 A (Qualcomm INC.) 12 May 1998 See the whole document	1,3,5

☐ Further documents are listed in the continuation of Box C.　　☒ See patent family annex.

* Special categories of cited documents:
"A" document defining the general state of the art which is not considered to be of particular relevence
"E" earlier application or patent but published on or after the international filing date
"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of citation or other special reason (as specified)
"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed
"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevence; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevence; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents,such combination being obvious to a person skilled in the art
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
24 OCTOBER 2000 (24.10.2000)	25 OCTOBER 2000 (25.10.2000)

Name and mailing address of the ISA/KR	Authorized officer
Korean Industrial Property Office Government Complex-Taejon, Dunsan-dong, So-ku, Taejon Metropolitan City 302-701, Republic of Korea	YOON, Byoung Soo
Facsimile No. 82-42-472-7140	Telephone No. 82-42-481-5709

INTERNATIONAL SEARCH REPORT

Information on patent family members

International application No.
PCT/KR00/00740

Patent document cited in search report	Publication date	Patent family member(s)	Publication date
US 5566206 A	10. 15. 1996	WO 9501032 A1	05.01.1995
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		EP 705512 B1	01.10.1997
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US 5751725 A	12. 05. 1998	EP 932963A1	04.08.1999
		CN 1234160 A	03.11.1999
		AU 4822097 A1	15.05.1998

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
20 September 2001 (20.09.2001)

PCT

(10) International Publication Number
WO 01/69801 A2

(51) International Patent Classification[7]: H04B

(21) International Application Number: PCT/US01/05576

(22) International Filing Date: 21 February 2001 (21.02.2001)

(25) Filing Language: English

(26) Publication Language: English

(30) Priority Data:
60/189,680 15 March 2000 (15.03.2000) US
60/207,700 26 May 2000 (26.05.2000) US

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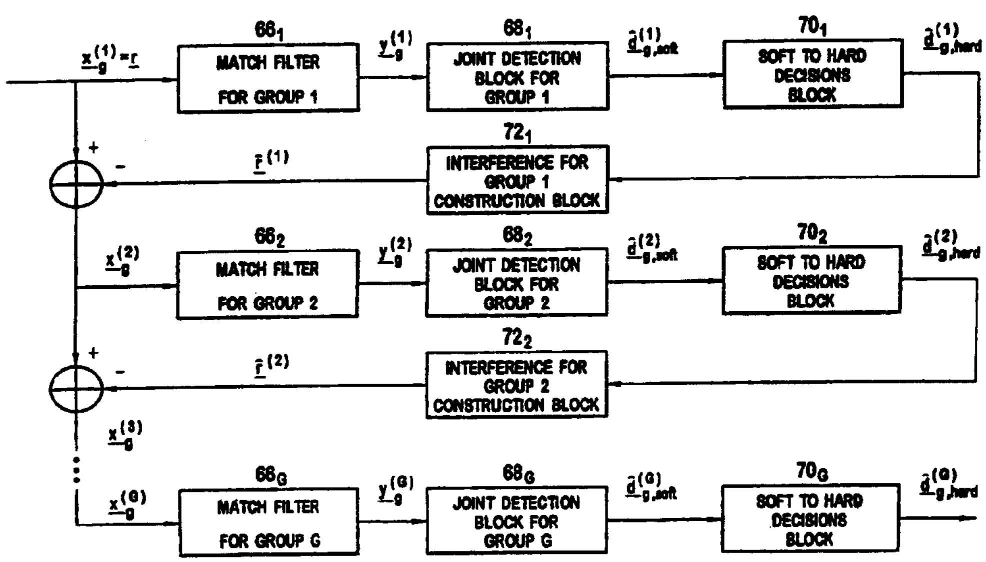
(81) Designated States *(national)*: CA, CN, JP, KR.

(84) Designated States *(regional)*: European patent (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE, TR).

Published:
— *without international search report and to be republished upon receipt of that report*

For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

(54) Title: MULTI-USER DETECTION USING AN ADAPTIVE COMBINATION OF JOINT DETECTION AND SUCCESSIVE INTERFERENCE CANCELLATION



(57) Abstract: A time division duplex communication system using code division multiple access transmits a plurality of data signals over a shared spectrum in a time slot. A combined signal is received over the shared spectrum in the time slot. The plurality of data signals are grouped into a plurality of groups. The combined signal is matched filtered based on in part symbol responses associated with the data signals of one of the groups. Data from each data signal in the one group is jointly detected. An interference signal is constructed based on in part the one group detected data. The constructed interference signal is subtracted from the combined signal. Data from the other groups is detected by processing the subtracted signal.

WO 01/69801 A2

MULTI-USER DETECTION USING AN ADAPTIVE COMBINATION OF JOINT DETECTION AND SUCCESSIVE INTERFERENCE CANCELLATION

This application claims priority to U.S. Provisional Patent Application No. 60/189,680, filed on March 15, 2000 and U.S. Provisional Patent Application No. 60/207,700, filed on May 26, 2000.

BACKGROUND

The invention generally relates to wireless communication systems. In particular, the invention relates to joint detection of multiple user signals in a wireless communication system.

Figure 1 is an illustration of a wireless communication system 10. The communication system 10 has base stations 12_1 to 12_5 which communicate with user equipments (UEs) 14_1 to 14_3. Each base station 12_1 has an associated operational area where it communicates with UEs 14_1 to 14_3 in its operational area.

In some communication systems, such as code division multiple access (CDMA) and time division duplex using code division multiple access (TDD/CDMA), multiple communications are sent over the same frequency spectrum. These communications are typically differentiated by their chip code sequences. To more efficiently use the frequency spectrum, TDD/CDMA communication systems use repeating frames divided into time slots for communication. A communication sent in such a system will have one or multiple associated chip codes and time slots assigned to it based on the communication's bandwidth.

Since multiple communications may be sent in the same frequency spectrum and at the same time, a receiver in such a system must distinguish between the multiple communications. One approach to detecting such signals is matched

filtering. In matched filtering, a communication sent with a single code is detected. Other communications are treated as interference. To detect multiple codes, a respective number of matched filters are used. Another approach is successive interference cancellation (SIC). In SIC, one communication is detected and the contribution of that communication is subtracted from the received signal for use in detecting the next communication.

In some situations, it is desirable to be able to detect multiple communications simultaneously in order to improve performance. Detecting multiple communications simultaneously is referred to as joint detection. Some joint detectors use Cholesky decomposition to perform a minimum mean square error (MMSE) detection and zero-forcing block equalizers (ZF-BLEs). These detectors have a high complexity requiring extensive receiver resources.

Accordingly, it is desirable to have alternate approaches to multi-user detection.

SUMMARY

A time division duplex communication system using code division multiple access transmits a plurality of data signals over a shared spectrum in a time slot. A combined signal is received over the shared spectrum in the time slot. The plurality of data signals are grouped into a plurality of groups. The combined signal is matched filtered based on in part symbol responses associated with the data signals of one of the groups. Data from each data signal in the one group is jointly detected. An interference signal is constructed based on in part the one group detected data. The constructed interference signal is subtracted from the combined signal. Data from the other groups is detected by processing the subtracted signal.

BRIEF DESCRIPTION OF THE DRAWING(S)

Figure 1 is a wireless communication system.

Figure 2 is a simplified transmitter and a receiver using joint detection.

Figure 3 is an illustration of a communication burst.

Figure 4 is a flow chart of adaptive combination of joint detection and successive interference cancellation.

Figure 5 is an illustration of an adaptive combination of joint detection and successive interference cancellation device.

Figures 6-12 are graphs comparing the performance of adaptive combination of joint detection and successive interference cancellation, full joint detection and a RAKE receiver.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENT(S)

Figure 2 illustrates a simplified transmitter 26 and receiver 28 using an adaptive combination of joint detection (JD) and successive interference cancellation (SIC), "SIC-JD", in a TDD/CDMA communication system. In a typical system, a transmitter 26 is in each UE 14_1 to 14_3 and multiple transmitting circuits 26 sending multiple communications are in each base station 12_1 to 12_5. A base station 12_1 will typically require at least one transmitting circuit 26 for each actively communicating UE 14_1 to 14_3. The SIC-JD receiver 28 may be at a base station 12_1, UEs 14_1 to 14_3 or both. The SIC-JD receiver 28 receives communications from multiple transmitters 26 or transmitting circuits 26.

Each transmitter 26 sends data over a wireless radio channel 30. A data generator 32 in the transmitter 26 generates data to be communicated over a reference channel to a receiver 28. Reference data is assigned to one or multiple codes and/or time slots based on the communications bandwidth requirements. A modulation and spreading device 34 spreads the reference data and makes the spread reference data time-multiplexed with a training sequence in the appropriate assigned time slots and codes. The resulting sequence is referred to as a communication burst. The communication burst is modulated by a modulator 36 to radio frequency. An antenna 38 radiates the RF signal through the wireless radio channel 30 to an antenna 40 of the receiver 28. The type of modulation used for the transmitted

communication can be any of those known to those skilled in the art, such as direct phase shift keying (DPSK) or quadrature phase shift keying (QPSK).

A typical communication burst 16 has a midamble 20, a guard period 18 and two data bursts 22, 24, as shown in Figure 3. The midamble 20 separates the two data bursts 22, 24 and the guard period 18 separates the communication bursts to allow for the difference in arrival times of bursts transmitted from different transmitters. The two data bursts 22, 24 contain the communication burst's data and are typically the same symbol length. The midamble contains a training sequence.

The antenna 40 of the receiver 28 receives various radio frequency signals. The received signals are demodulated by a demodulator 42 to produce a baseband signal. The baseband signal is processed, such as by a channel estimation device 44 and a SIC-JD device 46, in the time slots and with the appropriate codes assigned to the communication bursts of the corresponding transmitters 26. The channel estimation device 44 uses the training sequence component in the baseband signal to provide channel information, such as channel impulse responses. The channel information is used by the SIC-JD device 46 to estimate the transmitted data of the received communication bursts as hard symbols.

The SIC-JD device 46 uses the channel information provided by the channel estimation device 44 and the known spreading codes used by the transmitters 26 to estimate the data of the various received communication bursts. Although SIC-JD is described in conjunction with a TDD/CDMA communication system, the same approach is applicable to other communication systems, such as CDMA.

One approach to SIC-JD in a particular time slot in a TDD/CDMA communication system is illustrated in Figure 4. A number of communication bursts are superimposed on each other in the particular time slot, such as K communication bursts. The K bursts may be from K different transmitters. If certain transmitters are using multiple codes in the particular time slot, the K bursts may be from less than K transmitters.

Each data burst 22, 24 of the communication burst 16 has a predefined number of transmitted symbols, such as N_S. Each symbol is transmitted using a predetermined number of chips of the spreading code, which is the spreading factor (SF). In a typical TDD communication system, each base station 12_1 to 12_5 has an associated scrambling code mixed with its communicated data. The scrambling code distinguishes the base stations from one another. Typically, the scrambling code does not affect the spreading factor. Although the terms spreading code and factor are used hereafter, for systems using scrambling codes, the spreading code for the following is the combined scrambling and spreading codes. As a result, each data burst 22, 24 has N_S x SF chips. After passing through a channel having an impulse response of W chips, each received burst has a length of SF x N_s + W -1, which is also represented as N_c chips. The code for a k^{th} burst of the K bursts is represented by $C^{(k)}$.

Each k^{th} burst is received at the receiver and can be represented by Equation 1.

$$\underline{r}^{(k)} = A^{(k)}\underline{d}^{(k)},\ k = 1 \cdots K \qquad \text{Equation 1}$$

$\underline{r}^{(k)}$ is the received contribution of the k^{th} burst. $A^{(k)}$ is the combined channel response, being an N_c x N_s matrix. Each j^{th} column in $A^{(k)}$ is a zero-padded version of the symbol response $s^{(k)}$ of the j^{th} element of $d^{(k)}$. The symbol response $s^{(k)}$ is the convolution of the estimated response $\underline{h}^{(k)}$ and spreading code $C^{(k)}$ for the burst. $\underline{d}^{(k)}$ is the unknown data symbols transmitted in the burst. The estimated response for each k^{th} burst, $\underline{h}^{(k)}$, has a length W chips and can be represent by Equation 2.

$$\underline{h}^{(k)} = \gamma^{(k)} \cdot \underline{\tilde{h}}^{(k)} \qquad \text{Equation 2}$$

$\gamma^{(k)}$ reflects the transmitter gain and/or path loss. $\underline{\tilde{h}}^{(k)}$ represents the burst-specific fading channel response or for a group of bursts experiencing a similarly channel, $\underline{\tilde{h}}^{(g)}$ represents the group-specific channel response. For uplink communications, each $\underline{h}^{(k)}$ as well as each $\gamma^{(k)}$ and $\underline{\tilde{h}}^{(k)}$ are distinct. For the downlink, all of the bursts have the same $\underline{\tilde{h}}^{(k)}$ but each $\gamma^{(k)}$ is different. If transmit diversity is used in the downlink, each $\gamma^{(k)}$ and $\underline{\tilde{h}}^{(k)}$ are distinct.

The overall received vector from all K bursts sent over the wireless channel is per Equation 3.

$$\underline{r} = \sum_{i=1}^{K} \underline{r}^{(k)} + \underline{n}$$

Equation 3

$\underline{n}$ is a zero-mean noise vector.

By combining the $A^{(k)}$ for all data bursts into matrix A and all the unknown data for each burst $\underline{d}^{(k)}$ into matrix $\underline{d}$, Equation 1 becomes Equation 4.

$$\underline{r} = A\underline{d} + \underline{n}$$

Equation 4

SIC-JD determines the received power of each k^{th} burst. This determination may be based on apriori knowledge at the receiver 28, burst-specific channel estimation from a burst-specific training sequence, or a bank of matched filters. The K bursts are arranged in descending order based on the determined received power.

Bursts having roughly the same power level, such as within a certain threshold, are grouped together and are arranged into G groups, 48. The G groups are arranged into descending order by their power, such as from group 1 to G with

group 1 having the highest received power. Figure 5 is an illustration of a SIC-JD device 46 performing SIC-JD based on the G groups.

For the group with the highest received power, group 1, the symbol response matrix for only the bursts in group 1, $A_g^{(1)}$, is determined. $A_g^{(1)}$ contains only the symbol responses of the bursts in group 1. The received vector, $\underline{r}$, is modeled for group 1 as $\underline{x}_g^{(1)}$. As a result, Equation 4 becomes Equation 5 for group 1.

$$\underline{x}_g^{(1)} = A_g^{(1)} \underline{d}_g^{(1)} + \underline{n}$$

Equation 5

$\underline{d}_g^{(1)}$ is the data in the bursts of group 1. Equation 5 addresses both the effects of inter symbol interference (ISI) and multiple access interference (MAI). As a result, the effects of the other groups, groups 2 to G, are ignored.

The received vector, $\underline{x}_g^{(1)}$, is matched filtered to the symbol responses of the bursts in group 1 by a group 1 matched filter 66_1, such as per Equation 6, 50.

$$\underline{y}_g^{(1)} = A_g^{(1)H} \underline{x}_g^{(1)}$$

Equation 6

$\underline{y}_g^{(1)}$ is the matched filtered result.

A joint detection is performed on group 1 by a group 1 joint detection device 68_1 to make a soft decision estimate of $\hat{d}_{g,soft}^{(1)}$, using the matched filtered result $\underline{y}_g^{(1)}$. One JD approach is to compute the least-squares, zero-forcing, solution of Equation 7.

$$\hat{\underline{d}}_{g,soft}^{(1)} = \left(A_g^{(1)H} A_g^{(1)}\right)^{-1} \underline{y}_g^{(1)}$$

Equation 7

$A_g^{(1)H}$ is the hermetian of $A_g^{(1)}$. Another JD approach is to compute the minimum mean square error solution (MMSE) as per Equation 8.

$$\underline{\hat{d}}_{g,soft}^{(1)} = \left(A_g^{(1)^H} A_g^{(1)} + \sigma^2 I\right)^{-1} \underline{y}_g^{(1)}$$

Equation 8

I is the Identity matrix and σ^2 is the standard deviation.

One advantage to performing joint detection on only a group of bursts is that the complexity of analyzing a single group versus all the signals is reduced. Since $A_g^{(1)^H}$ and $A_g^{(1)}$ are banded block Toeplitz matrices, the complexity in solving either Equation 7 or 8 is reduced. Additionally, Cholesky decomposition may be employed with a negligible loss in performance. Cholesky decomposition performed on a large number of bursts is extremely complex. However, on a smaller group of users, Cholesky decomposition can be performed at a more reasonable complexity.

The soft decisions, $\hat{d}_{g,soft}^{(1)}$, are converted into hard decisions, $\hat{d}_{g,hard}^{(1)}$, by soft to hard decision block 70_1 as the received data for group 1, 54. To process the other weaker groups, the multiple access interference caused by group 1 onto the weaker groups is estimated by a group 1 interference construction block 72_1 using Equation 9, 56.

$$\underline{\hat{r}}^{(1)} = A_g^{(1)} \underline{\hat{d}}_{g,hard}^{(1)}$$

Equation 9

$\underline{\hat{r}}^{(1)}$ is the estimated contribution of group 1 to $\underline{r}$.

For the next group 2, the estimated contribution of group 1 is removed from the received vector, $\underline{x}_g^{(1)}$, to produce $\underline{x}_g^{(2)}$, such as by a subtractor 74_1, as per Equation 10, 58.

$$\underline{x}_g^{(2)} = \underline{x}_g^{(1)} - \underline{\hat{r}}^{(1)}$$

Equation 10

As a result, multiple access interference from group 1 is effectively canceled from the received signal. The next strongest group, group 2, is processed similarly using $\underline{x}_g^{(2)}$, with group 2 matched filter 66_2, group 2 JD block 68_2, soft to hard decision block 70_2 and group 2 interference construction block 72_2, 60. The constructed group 2 interference, $\hat{\underline{r}}^{(2)}$, is subtracted, such as by subtractor 24_2, from the interference cancclled signal for group 2, $\underline{x}_g^{(2)} - \hat{\underline{r}}^{(2)} = \underline{x}_g^{(3)}$, 62. Using this procedure, each group is successively processed until the final group G. Since group G is the last group, the interference construction does not need to be performed. Accordingly, group G is only processed with group G matched filter 66_G, group G JD block 68_G and soft to hard decisions block 70_G to recovery the hard symbols, 64.

When SIC-JD is performed at a UE 14_1, it may not be necessary to process all of the groups. If all of the bursts that the UE 14_1 is intended to receive are in the highest received power group or in higher received power groups, the UE 14_1 will only have to process the groups having its bursts. As a result, the processing required at the UE 14_1 can be further reduced. Reduced processing at the UE 14_1 results in reduced power consumption and extended battery life.

SIC-JD is less complex than a single-step JD due to the dimension N_c x $K \cdot N_s$ matrix being replaced with G JD stages of dimension N_c x $n_i \cdot N_s$, where i = 1 to G. n_i is the number of bursts in the i^{th} group. The complexity of JD is proportional to the square to cube of the number of bursts being jointly detected.

An advantage of this approach is that a trade-off between computational complexity and performance can be achieved. If all of the bursts are placed in a single group, the solution reduces to a JD problem. The singlc grouping can be achieved by either forcing all the bursts into one group or using a broad threshold. Alternately, if the groups contain only one signal or only one signal is received, the solution reduces to a SIC-LSE problem. Such a situation could result using a narrow threshold or forcing each burst into its own group, by hard limiting the group size.

By selecting the thresholds, an optional tradeoff between performance and complexity can be achieved.

Figures 6 to 12 are simulation results that compare the bit error rate (BER) performance of SIC-JD to full JD and RAKE-like receivers under various multi-path fading channel conditions. The parameters chosen are those of the 3G UTRA TDD CDMA system: SF = 61 and W = 57. Each TDD burst/time-slot is 2560 chips or 667 microseconds long. The bursts carry two data fields with N_S QPSK symbols each, a midamble field and a guard period. Each simulation is run over 1000 timeslots. In all cases the number of bursts, K is chosen to be 8. All receivers are assumed to have exact knowledge of the channel response of each burst, which is used to perfectly rank and group the bursts. The channel response is assumed to be time-invariant over a time-slot, but successive time-slots experience uncorrelated channel responses. No channel coding was applied in the simulation. The JD algorithm jointly detects all K bursts. The RAKE-like receiver was a bank of matched filters, $\hat{\underline{d}}^{(i)} = A^{(i)^H} \underline{r}^{(i)}$, for an i^{th} burst's code. The maximal ratio combiner (MRC) stage is implicit in these filters because they are matched to the entire symbol-response.

The performance was simulated under fading channels with multi-path profiles defined by the ITU channel models, such as the Indoor A, Pedestrian A, Vehicular A models, and the 3GPP UTRA TDD Working Group 4 Case 1, Case 2 and Case 3 models. In Vehicular A and Case 2 channels, the SIC-JD suffered a degradation of up to 1 decibel (dB) as compared to the full JD in the 1% to 10% BER range. For all other channels, the SIC-JD performance was within 0.5 dB of that of the full JD. Since Vehicular A and Case 2 represent the worst-case amongst all cases studied, only the performance plots are shown. Amongst all channels simulated, Vehicular A and Case 2 have the largest delay spread. Vehicular A is a six tap model with relative delays of 0, 310, 710, 1090, 1730 and 2510 nanoseconds and relative average powers of 0, -1, -9, -10, -15 and -20 decibels (dB). Case 2 is a

3 tap model, all with the same average power and with relative delays of 0, 976 and 1200 nanoseconds.

Figures 6 and 7 compare the bit error rate (BER) vs. the chip-level signal to noise ratio (SNR) performance of the SIC-LSE receiver with the full JD and RAKE-like receivers under two multi-path fading channel conditions. The group size is forced to be 1, to form K groups, both, at the transmitter and receiver. The theoretical binary phase shift keying (BPSK) BER in an additive white gaussian noise (AWGN) channel that provides a lower bound to the BER is also shown. The BER is averaged over all bursts. Figure 6 represents the distinct channel case wherein each burst is assumed to pass through an independently fading channel but all channels have the same average power leading to the same average SNR. Thus, in this case, $\underline{\tilde{h}}^{(i)}, i = 1 \cdots K$ are distinct while $\gamma^{(i)}, i = 1 \cdots K$ are all equal. Such a situation exists in the uplink where the power control compensates for long-term fading and/or path-loss but not for short-term fading. At each time-slot, the bursts were arranged in power based upon the associated $\underline{h}^{(i)}, i = 1 \cdots K$. Figure 7 shows similar plots for the common channel case. All bursts are assumed to pass through the same multi-path channel, i.e., $\underline{\tilde{h}}^{(i)}, i = 1 \cdots K$ and are all equal, but with different $\gamma^{(i)}, i = 1 \cdots K$. The $\delta^{(i)}$ are chosen such that neighboring bursts have a power separation of 2 dB when arranged by power level. Such difference in power can exist, for instance, in the downlink where the base station 12_1 applies different transmit gains to bursts targeted for different UEs 14_1 to 14_3. Figures 6 and 7 show that in the range of 1% to 10% bit error rate (BER), the SIC-LSE suffers a degradation of less than 1 dB as compared to the JD. This is often the range of interest for the uncoded BER (raw BER). The RAKE receiver exhibits significant degradation, since it does not optimally handle the ISI. As the power differential between bursts increases, the performance of SIC-LSE improves. Depending upon

the channel, a power separation of 1 to 2 dB is sufficient to achieve a performance comparable to that of the full JD.

Figures 8, 9, 10 and 11 compare the BER vs. SNR performance of the SIC-JD receiver with the full JD and RAKE-like receivers under two multi-path fading channels. The 8 codes are divided into 4 groups of 2 codes each at the transmitter and receiver. The BER is averaged over all bursts. Figures 8 and 9 represent the distinct channel case wherein different groups are assumed to pass through independently fading channels. However, all channels have the same average power leading to the same average SNR. All bursts within the same group are subjected to an identical channel response. In this case, $\underline{\tilde{h}}_g^{(g)}, g = 1 \cdots G$ are all distinct, but the channel responses, $h_g^{(i)}, i = 1, \cdots n_g$, for each burst in the group are equal. n_g is the number of bursts in the g^{th} group. This potentially represents a multi-code scenario on the uplink, where each UE 14_1 transmits two codes. The SIC-JD receiver 28 groups the multi-codes associated with a single UE 14_1 into the same group, thus forming 4 groups. Figures 10 and 11 represent the common channel case. All groups are assumed to pass through the same multi-path channel, i.e., $\underline{\tilde{h}}_g^{(i)}, g = 1 \cdots n_g$ are all equal, but with different $\gamma_g, g = 1 \cdots G$. The γ_g are chosen such that, when arranged according to power, neighboring groups have a power separation of 2 dB. This potentially represents a multi-code scenario on the downlink where the base station 12_1 transmits 2 codes per UE 14_1. Figures 10 and 11 show a trend similar to that observed for the SIC-LSE shown in Figures 8 and 9. SIC-JD has a performance comparable (within a dB) to the JD in the region of 1% to 10% BER, which is the operating region of interest for the uncoded BER. Depending upon the channel, a power separation of 1 to 2 dB is sufficient to achieve a performance of SIC-LSE comparable to that of the full JD. As shown, performance improves as the power separation between bursts increases.

Figure 12 is similar to Figure 10, except that there are only two groups with 4 bursts each. As shown in Figure 12, SIC-JD has a performance comparable (within a dB) to JD in the region of 1% to 10% BER.

The complexity of SIC-JD is less than full JD. The reduced complexity stems from the replacement of a single-step JD which is a dimension $N_c \times K \cdot N_s$ with G JD stages of dimension $N_c \times n_i \cdot N_s$, $i = 1 \cdots G$. Since, typically, JD involves a matrix inversion, whose complexity varies as the cube of the number of bursts, the overall complexity of the multi-stage JD can be significantly lower than that of the single-stage full JD. Furthermore, the complexity of the SIC part varies only linearly with the number of bursts, hence it does not offset this complexity advantage significantly. For instance, the complexity of the G-1 stages of interference cancellation can be derived as follows. Since successive column blocks of $A_g^{(i)}$ are shifted versions of the first block and assuming that elements of $\underline{\hat{d}}_{g,hard}^{(i)}$ belong to 1 of 4 QPSK constellation points, the $4 \cdot n_i$ possible vectors can be computed that are needed in computing the product $A_g^{(i)} \underline{\hat{d}}_{g,hard}^{(i)}$. This step requires $4\alpha \cdot (SF + W - 1) \cdot \frac{Rate}{10^6} \sum_{i=1}^{G-1} n_i$ million real operations per sec (MROPS). $\alpha = 4$ is the number of real operations per complex multiplication or multiply and accumulate (MAC). *Rate* is the number of times the SIC-JD is performed per second. With these $4 \cdot n_i$ vectors already computed, the computation of $\underline{x}_g^{(i+1)}$ requires $\frac{\alpha}{2} \cdot N_s \cdot (SF + W - 1) \cdot \frac{Rate}{10^6} \sum_{i=1}^{G-1} n_i$ MROPS. The factor of $\frac{\alpha}{2}$ comes from the fact that only complex additions are involved. Hence, only 2 real operations are required for

each complex operation. It then follows that the complexity of $G-1$ stages of interference cancellation can be expressed by Equation 11.

$$Z=\alpha(SF+W-1)\cdot\left(4+\frac{N_s}{2}\right)\cdot\frac{Rate}{10^6}\cdot\sum_{i=1}^{G-1}n_i$$

Equation 11

The complexity of converting soft to hard decisions is negligible.

There are several well-known techniques to solve the matrix inversion of JD. To illustrate the complexity, an approach using a very efficient approximate Cholesky factor algorithm with negligible loss in performance as compared to the exact Cholesky factor algorithm was used. The same algorithm can be employed to solve group-wise JD. The complexity of the full JD and the SIC-JD for the 3GPP UTRA TDD system is shown in Table 1. Table 1 compares their complexity for various group sizes. It can be seen that as K increases or as the group size decreases the complexity advantage of the SIC-JD over the full JD increases. The complexity for group size 1, of the SIC-LSE, varies linearly with K and is 33% that of the full JD for $K = 16$. Note that maximum number of bursts in the UTRA TDD system is 16. The complexity advantage of the SIC-JD over full JD will be even more pronounced when the exact Cholesky decomposition is employed. Exact Cholesky decomposition's complexity exhibits a stronger dependence on K, leading to more savings as the dimension of the JD is reduced via SIC-JD.

Total number of bursts	Complexity of the SIC-JD expressed as a percentage of the complexity of the single-step JD of all K bursts			
	K groups of size 1 each (SIC-LSE)	K / 2 groups of size 2 each	K / 4 groups of size 4 each	K / 8 groups of size 8 each
8	63 %	67 %	76 %	100 %
16	33 %	36 %	41 %	57 %

Table 1

As shown in Table 1, when the number and size of codes is made completely adaptive on an observation interval-by-observation interval basis, the SIC-JD provides savings, on average, over full JD. Since, on average, all bursts do not arrive at the receiver with equal power, depending upon the grouping threshold, the size of the groups will be less then the total number of arriving bursts. In addition, a reduction in peak complexity is also possible if the maximum allowed group size is hard-limited to be less than the maximum possible number of bursts. Such a scheme leads to some degradation in performance when the number of bursts arriving at the receiver with the roughly the same power exceeds the maximum allowed group size. Accordingly, SIC-JD provides a mechanism to trade-off performance with peak complexity or required peak processing power.

* * *

CLAIMS

What is claimed is:

1. A method for use in receiving a plurality of data signals transmitted over a shared spectrum in a time slot in a time division duplex communication system using code division multiple access, the method comprising:

receiving a combined signal over the shared spectrum in the time slot;

grouping the plurality of data signals into a plurality of groups;

match filtering the combined signal based on in part symbol responses associated with the data signals of one of the groups;

jointly detecting data from each data signal in the one group;

constructing an interference signal based on in part the one group detected data;

subtracting the constructed interference signal from the combined signal; and

determining data from a group other than the one group by processing the subtracted signal.

2. The method of claim 1 wherein the jointly detecting is performed using least squares estimation.

3. The method of claim 1 wherein the jointly detecting is performed using minimum mean square error estimation.

4. A method for use in receiving a plurality of data signals transmitted over a shared spectrum in a time slot in a time division duplex communication system using code division multiple access, the method comprising:

(a) receiving a combined signal as an input signal over the shared spectrum in the time slot;

(b) grouping the plurality of data signals into a plurality of groups, at least one of the groups having a plurality of data signals;

(c) match filtering the input signal based on in part symbol responses associated with each data signal of a first group of the groups;

(d) jointly detecting data from each data signal in the first group;

(e) constructing an interference signal based on in part the first group detected data;

(f) subtracting the constructed interference signal from the input signal as an input signal for subsequent processing;

(g) match filtering the subtracted signal based on in part symbol responses associated with the data signal of a subsequent group of the groups;

(h) jointly detecting data from each data signal in the subsequent group; and

(i) successively repeating steps (e) through (h) for remaining groups of the plurality of groups where, for each remaining group, the subsequent group acts as the first group for that remaining group and that remaining group acts as the subsequent group.

5. A method for use in a receiver for receiving a plurality of data signals transmitted over a shared spectrum in a time slot in a time division duplex communication system using code division multiple access, the method comprising:

receiving a combined signal over the shared spectrum in the time slot;

estimating a received power level for each data signal;

selectively grouping data signals of the plurality of data signals based on in part the received power level of the data signals into at least one group; and

separately detecting data within each group for that group's data signals.

6. The method of claim 5 wherein the estimating the received power level for each data signal is based on in part apriori knowledge at the receiver.

7. The method of claim 5 wherein the estimating the received power level for each data signal is based on in part a power level of a training sequence associated with each data signal.

8. The method of claim 5 wherein the estimating the received power level for each data signal is performed using a bank of matched filters, each matched filter matched to a code of a respective one of the data signals.

9. The method of claim 5 wherein the selectively grouping data signals groups data signals within a certain threshold power level into a group.

10. The method of claim 9 wherein the certain threshold power level is one decibel.

11. The method of claim 9 wherein the certain threshold is adjusted to achieve a desired bit error rate at the receiver.

12. The method of claim 5 further comprising forcing all of the data signals into a single group to override the step of selectively grouping.

13. The method of claim 5 further comprising forcibly grouping each data signal into its own group to override the step of selectively grouping.

14. A method for use in a receiver for adjusting a trade-off between complexity and performance in detecting data from data signals transmitted over a shared spectrum in a time slot in a time division duplex communication system using code division multiple access, the method comprising:

grouping the data signals into at least one group; wherein to reduce the complexity, increasing a number of data signal groups, and to increase the performance, decreasing a number of data signal groups; and

jointly detecting data in each group.

15. The method of claim 14 further comprising:

determining a received power of each data signal; wherein the grouping is performed so that all data signals within each group are within a certain threshold power level and to reduce complexity, the certain threshold is increased and to increase performance, the certain threshold is reduced.

16. The method of claim 14 wherein to reduce the complexity, each group contains one of the data signals.

17. The method of claim 14 wherein to increase the performance, the at least one group is a single group.

18. A receiver for use in a time division duplex communication system using code division multiple access, the system communicating using multiple communication bursts in a time slot, the receiver comprising:

an antenna for receiving radio frequency signals including the multiple communication bursts;

a demodulator for demodulating radio frequency signals to produce a baseband signal;

a channel estimation device for estimating a channel response for the bursts;

a successive interference cancellation joint detection (SIC-JD) device comprising:

a first joint detection block for detecting data within the baseband signal for a first group of bursts of the multiple bursts;

a first interference construction block for constructing an estimate of interference of the first group bursts;

a subtractor for subtracting the first group interference from the baseband signal; and

a second joint detection block for detecting data within the subtracted signal for a second group of bursts of the multiple bursts.

19. The receiver of claim 18 wherein the SIC-JD device further comprises:

a plurality of additional joint detection blocks for detecting data for additional groups of bursts of the multiple bursts.

20. The receiver of claim 18 wherein the SIC-JD device further comprises:

a first matched filter for processing the baseband signal to match symbol responses of the data signals in the first group; and

a second matched filter for processing the subtracted signal to match symbol responses of the data signals in the second group.

21. The receiver of claim 18 wherein an output of the first and second joint detection blocks are soft symbols, the SIC-JD device further comprising a first and second soft to hard decision block for converting the first and second joint detection block outputs into hard symbols.

22. A device for use in a receiver of a time division duplex communication system using code division multiple access, the system communicating using multiple communication bursts in a time slot, the device comprising:

an input configured to receive a baseband signal associated with received bursts within a time slot;

a first joint detection block for detecting data within the baseband signal for a first group of bursts of the received bursts;

a first interference construction block for constructing an estimate of interference of the first group bursts;

a subtractor for subtracting the first group interference from the baseband signal; and

a second joint detection block for detecting data within the subtracted signal for a second group of bursts of the received bursts.

23. The device of claim 22 further comprising additional joint detection blocks for detecting data for additional groups of bursts of the multiple bursts.

24. The device of claim 22 further comprising:

a first matched filter for processing the baseband signal to match symbol responses of the received bursts of the first group; and

a second matched filter for processing the subtracted signal to match symbol responses of the received bursts of the second group.

25. The device of claim 22 wherein an output of the first and second joint detection blocks are soft symbols, the device further comprising a first and second soft to hard decision block converting the first and second joint detection block outputs into hard symbols.

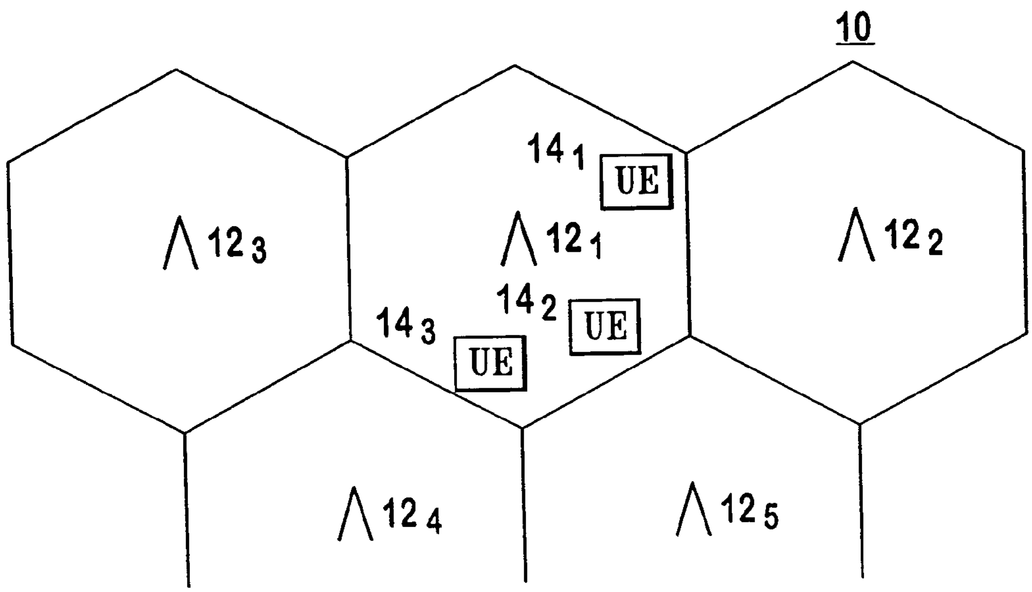

10
14_1
UE
12_3
12_1
12_2
14_2
UE
14_3
UE
12_4
12_5

FIG. 1

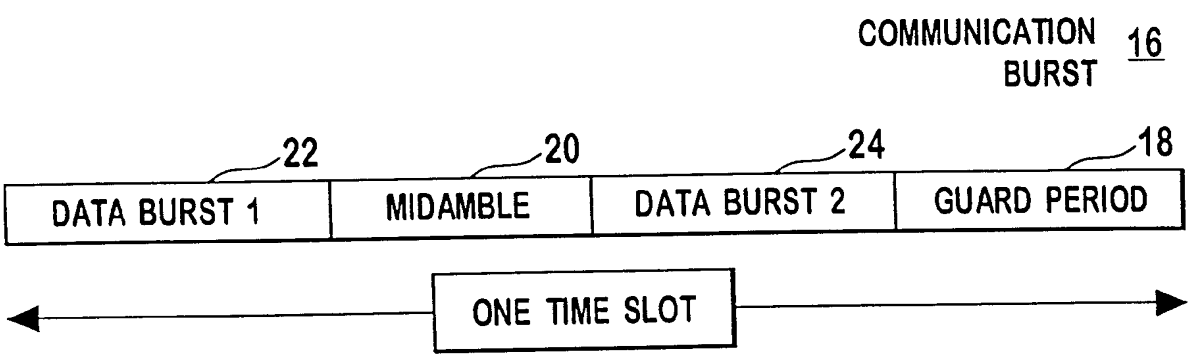

COMMUNICATION BURST
16
22
20
24
18
DATA BURST 1
MIDAMBLE
DATA BURST 2
GUARD PERIOD
ONE TIME SLOT

FIG. 3

2/8

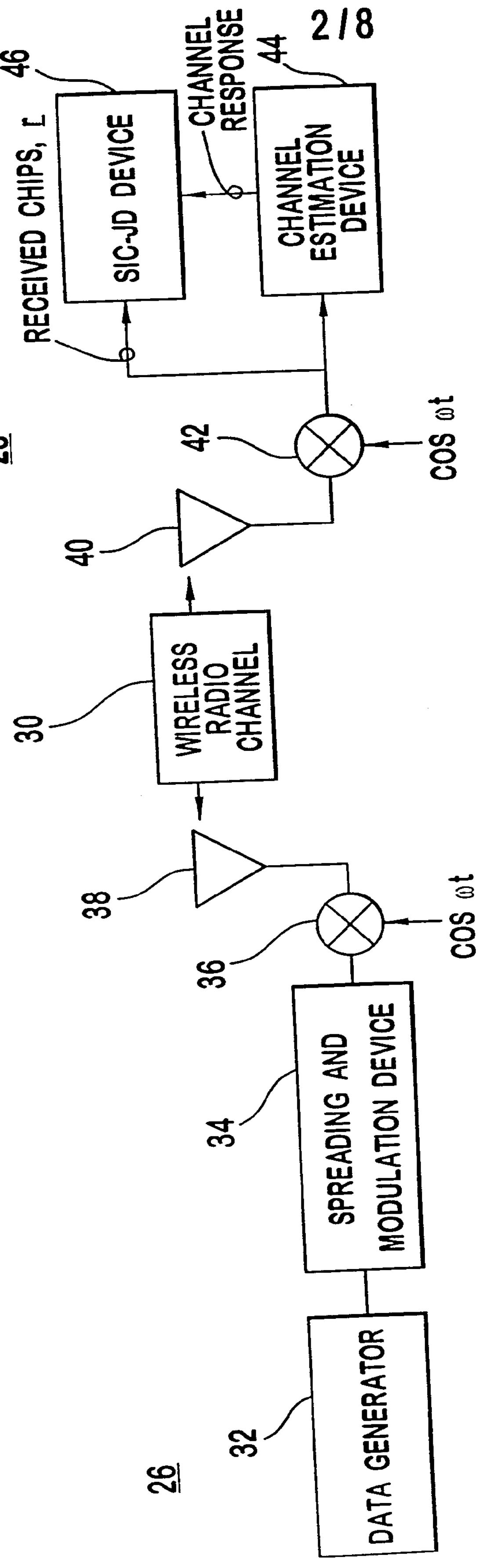

26
DATA GENERATOR
32
SPREADING AND MODULATION DEVICE
34
36
COS ωt
38
WIRELESS RADIO CHANNEL
30
40
42
COS ωt
28
RECEIVED CHIPS, r
SIC-JD DEVICE
46
CHANNEL RESPONSE
CHANNEL ESTIMATION DEVICE
44

FIG. 2

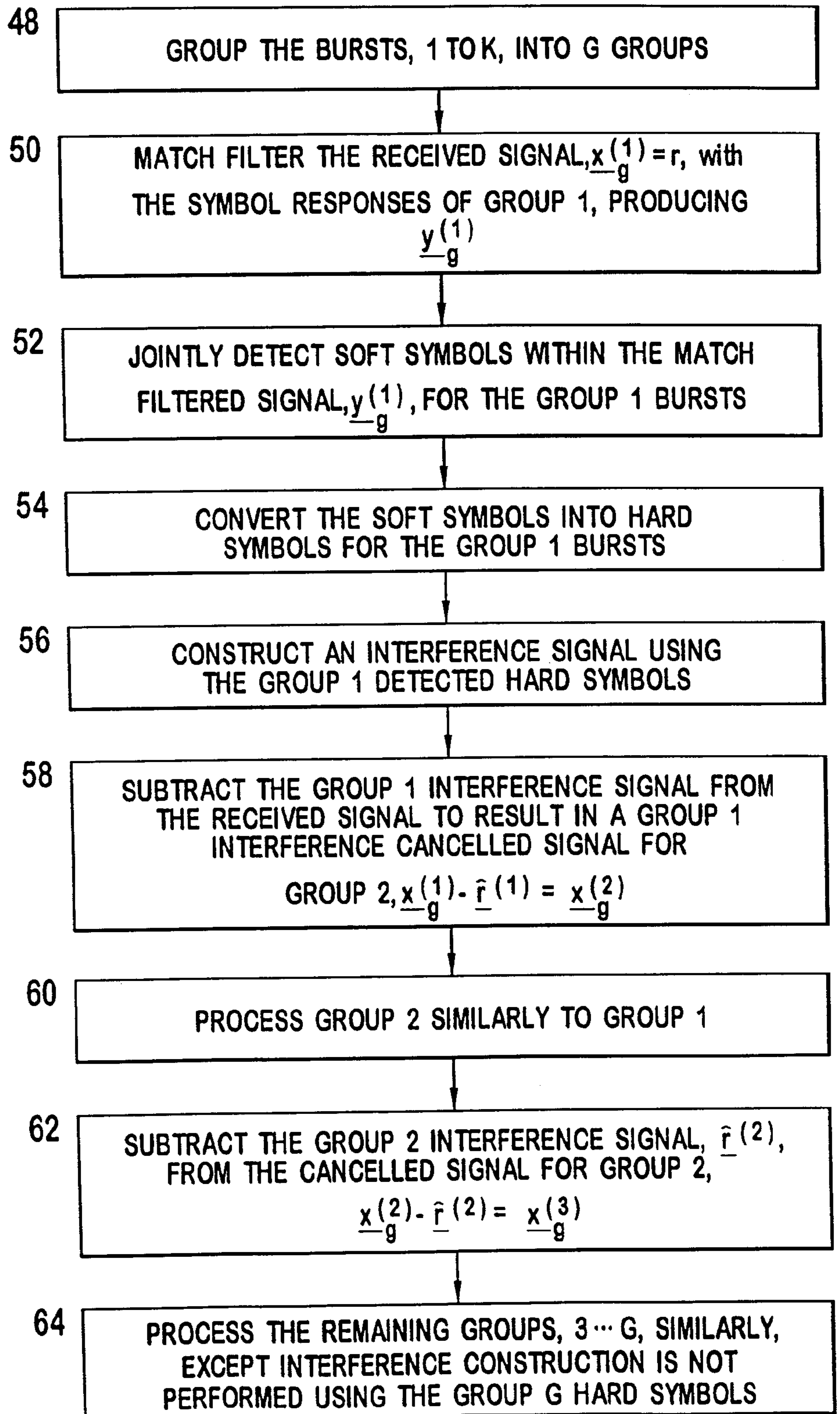

48
GROUP THE BURSTS, 1 TO K, INTO G GROUPS
50
MATCH FILTER THE RECEIVED SIGNAL, $\underline{x}_g^{(1)} = r$, with THE SYMBOL RESPONSES OF GROUP 1, PRODUCING $\underline{y}_g^{(1)}$
52
JOINTLY DETECT SOFT SYMBOLS WITHIN THE MATCH FILTERED SIGNAL, $\underline{y}_g^{(1)}$, FOR THE GROUP 1 BURSTS
54
CONVERT THE SOFT SYMBOLS INTO HARD SYMBOLS FOR THE GROUP 1 BURSTS
56
CONSTRUCT AN INTERFERENCE SIGNAL USING THE GROUP 1 DETECTED HARD SYMBOLS
58
SUBTRACT THE GROUP 1 INTERFERENCE SIGNAL FROM THE RECEIVED SIGNAL TO RESULT IN A GROUP 1 INTERFERENCE CANCELLED SIGNAL FOR GROUP 2, $\underline{x}_g^{(1)} - \hat{\underline{r}}^{(1)} = \underline{x}_g^{(2)}$
60
PROCESS GROUP 2 SIMILARLY TO GROUP 1
62
SUBTRACT THE GROUP 2 INTERFERENCE SIGNAL, $\hat{\underline{r}}^{(2)}$, FROM THE CANCELLED SIGNAL FOR GROUP 2, $\underline{x}_g^{(2)} - \hat{\underline{r}}^{(2)} = \underline{x}_g^{(3)}$
64
PROCESS THE REMAINING GROUPS, 3 ··· G, SIMILARLY, EXCEPT INTERFERENCE CONSTRUCTION IS NOT PERFORMED USING THE GROUP G HARD SYMBOLS

FIG. 4

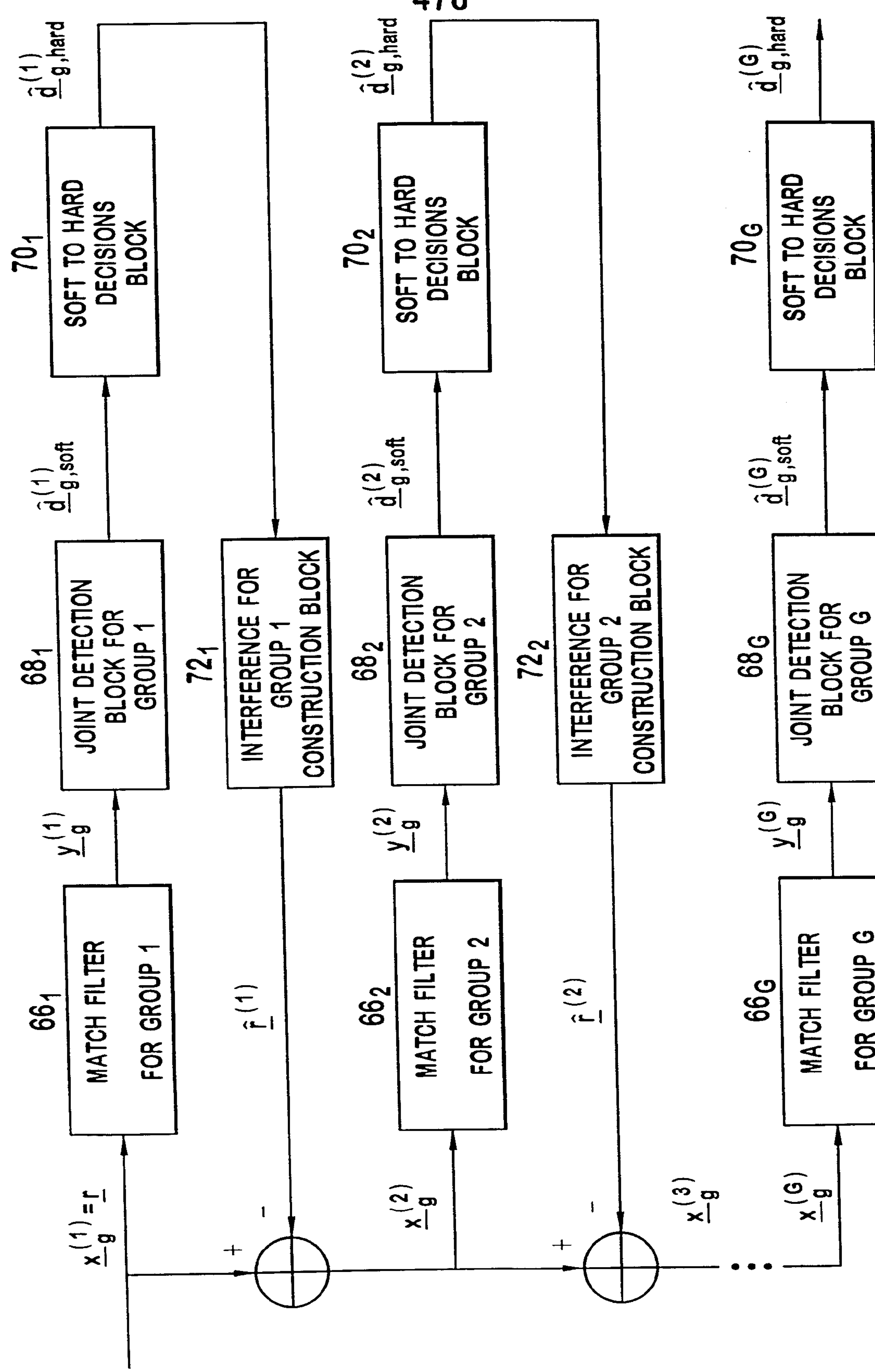

$x_g^{(1)} = r$
66_1
MATCH FILTER FOR GROUP 1
$y_g^{(1)}$
68_1
JOINT DETECTION BLOCK FOR GROUP 1
$\hat{d}_{g,soft}^{(1)}$
70_1
SOFT TO HARD DECISIONS BLOCK
$\hat{d}_{g,hard}^{(1)}$
72_1
INTERFERENCE FOR GROUP 1 CONSTRUCTION BLOCK
$\hat{r}^{(1)}$
+
-
$x_g^{(2)}$
66_2
MATCH FILTER FOR GROUP 2
$y_g^{(2)}$
68_2
JOINT DETECTION BLOCK FOR GROUP 2
$\hat{d}_{g,soft}^{(2)}$
70_2
SOFT TO HARD DECISIONS BLOCK
$\hat{d}_{g,hard}^{(2)}$
72_2
INTERFERENCE FOR GROUP 2 CONSTRUCTION BLOCK
$\hat{r}^{(2)}$
+
-
$x_g^{(3)}$
$x_g^{(G)}$
66_G
MATCH FILTER FOR GROUP G
$y_g^{(G)}$
68_G
JOINT DETECTION BLOCK FOR GROUP G
$\hat{d}_{g,soft}^{(G)}$
70_G
SOFT TO HARD DECISIONS BLOCK
$\hat{d}_{g,hard}^{(G)}$

FIG. 5

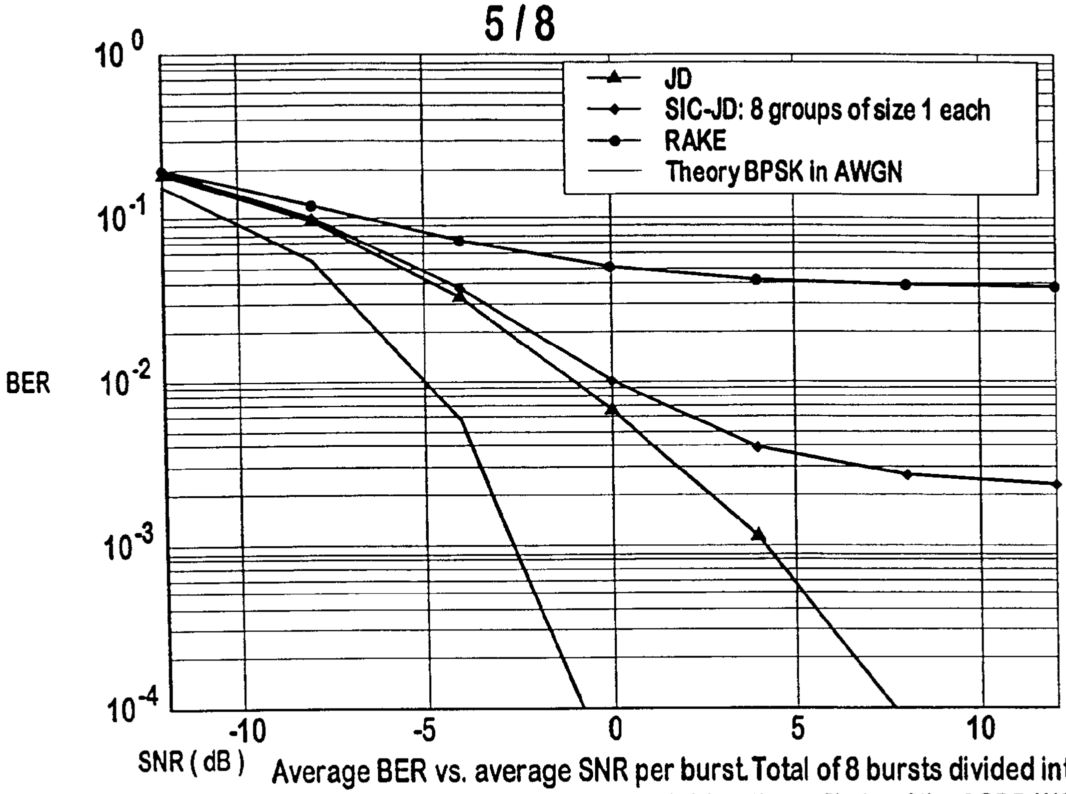


FIG. 6

Average BER vs. average SNR per burst. Total of 8 bursts divided into 8 groups with 1 burst per group. Multi-path profile is of the 3GPP WG4 Case 2 type. All 8 bursts have the same average SNR but pass through uncorrelated channel.

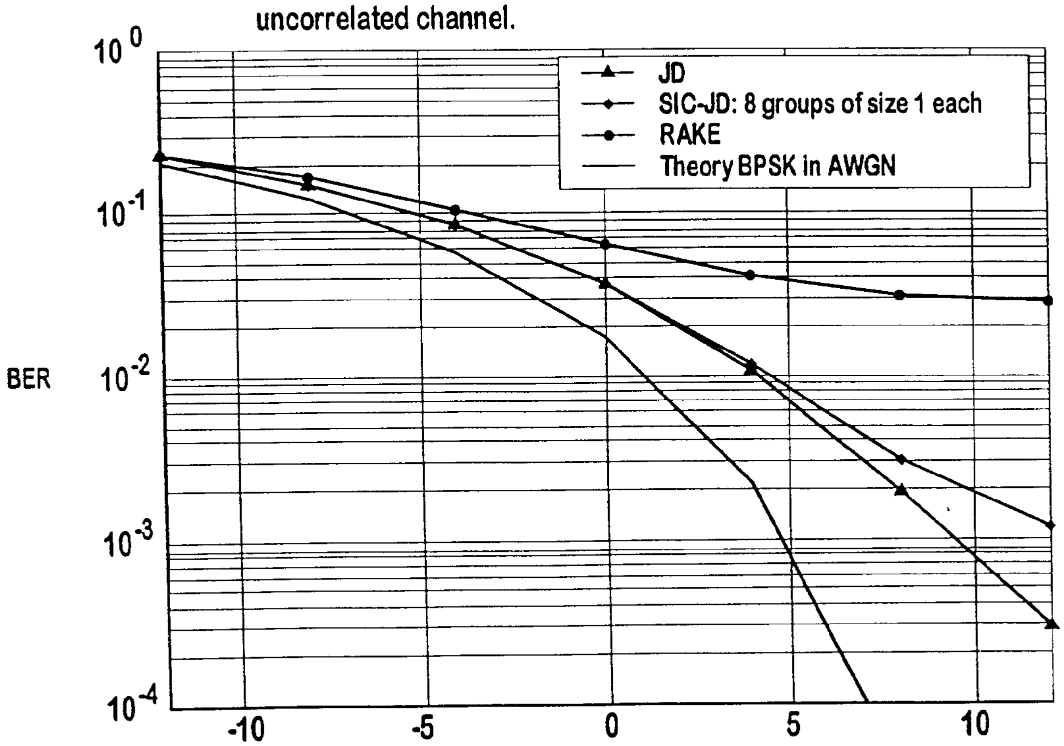


FIG. 7

Average BER vs. average SNR per burst. Total of 8 bursts divided into 8 groups with 1 burst per group. Multi-path profile is of the 3GPP WG4 Case 2 type. All 8 bursts pass through a common channel but their average SNR is separated by 2 dB.

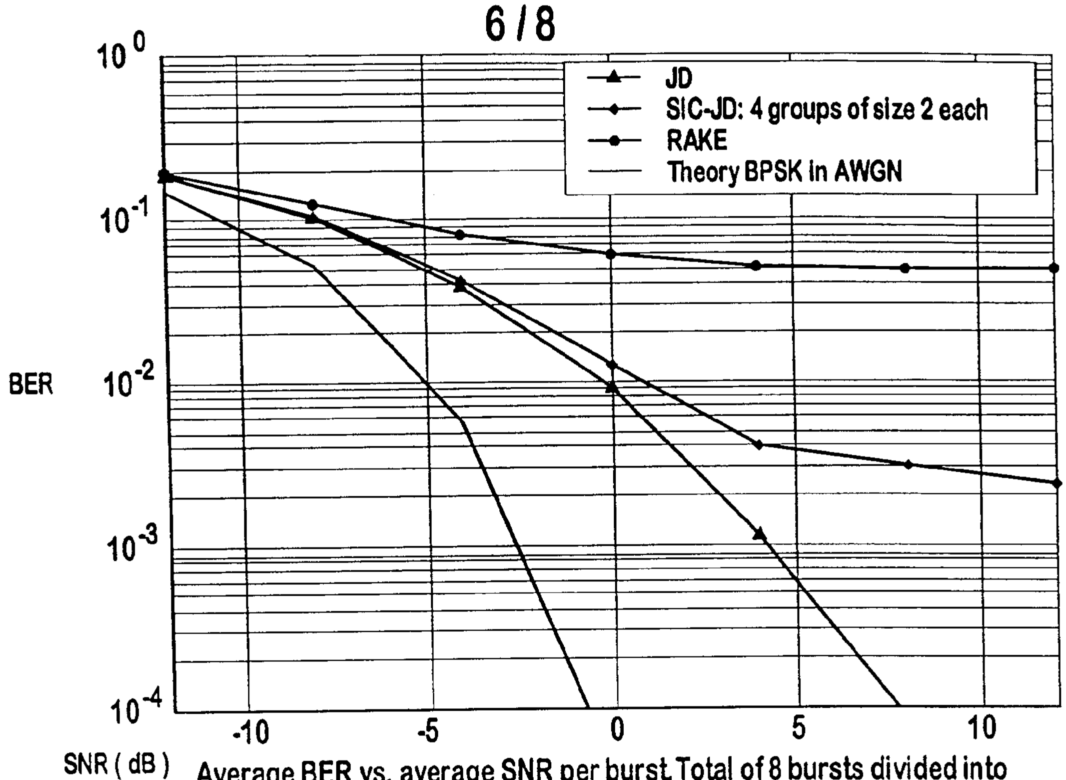


FIG. 8

Average BER vs. average SNR per burst. Total of 8 bursts divided into 4 groups with 2 bursts per group. All bursts in the same group are subjected to the same channel. All 4 groups have the same average SNR but pass through uncorrelated channels with multi-path profile of type ITU Vehicular A.

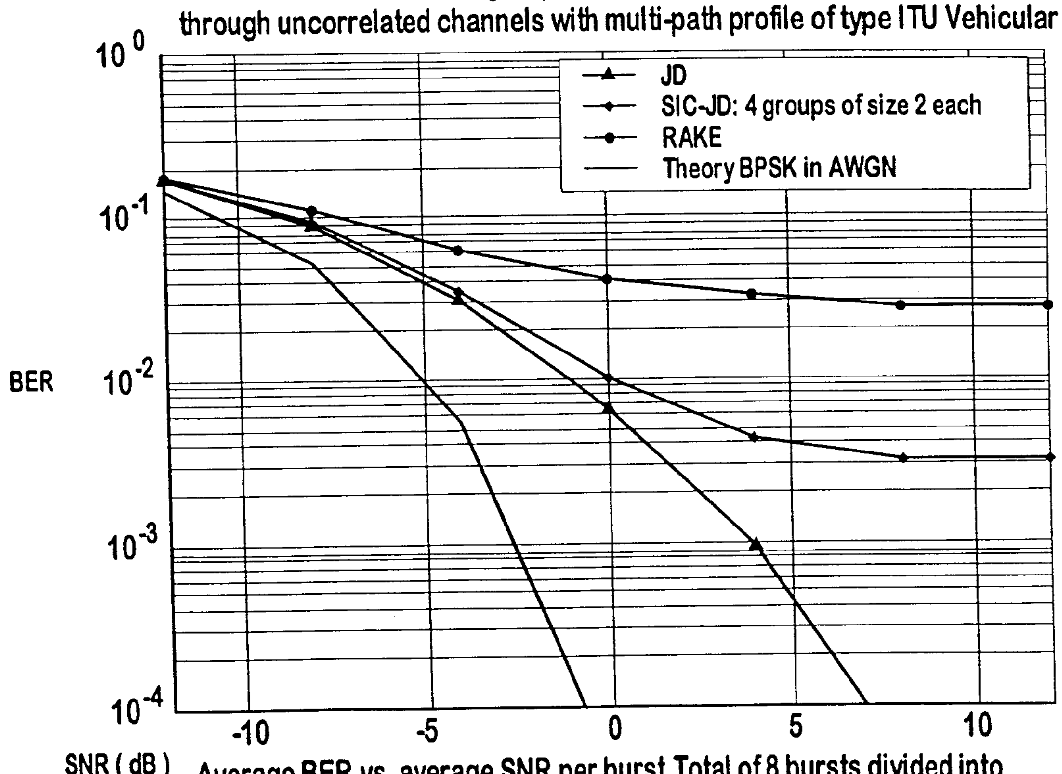


FIG. 9

Average BER vs. average SNR per burst. Total of 8 bursts divided into 4 groups with 2 bursts per group. All bursts in the same group are subjected to the same channel. All 4 groups have the same average SNR but pass through uncorrelated channels with multi-path profile of type 3GPP WG4 Case 2.

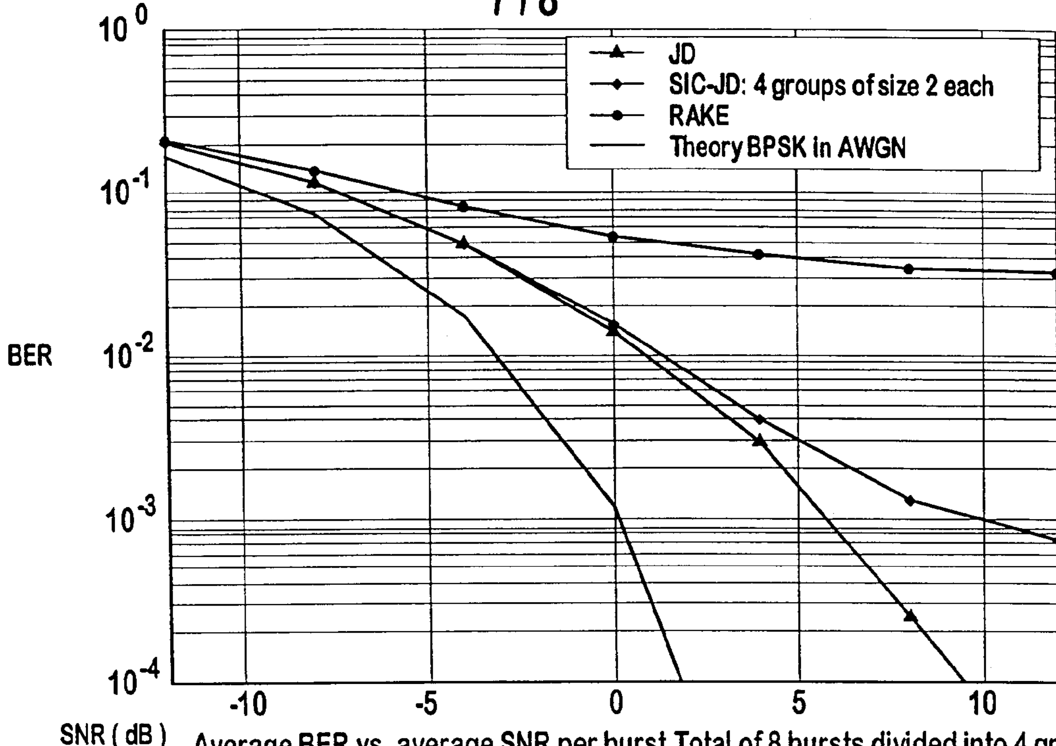


FIG. 10

Average BER vs. average SNR per burst. Total of 8 bursts divided into 4 groups with 2 bursts per group. All bursts in the same group are subjected to the same channel. All 4 groups pass through a common channel but their average SNR is seperated by 2 dB. Multi-path profile is of type ITU Vehicular A.

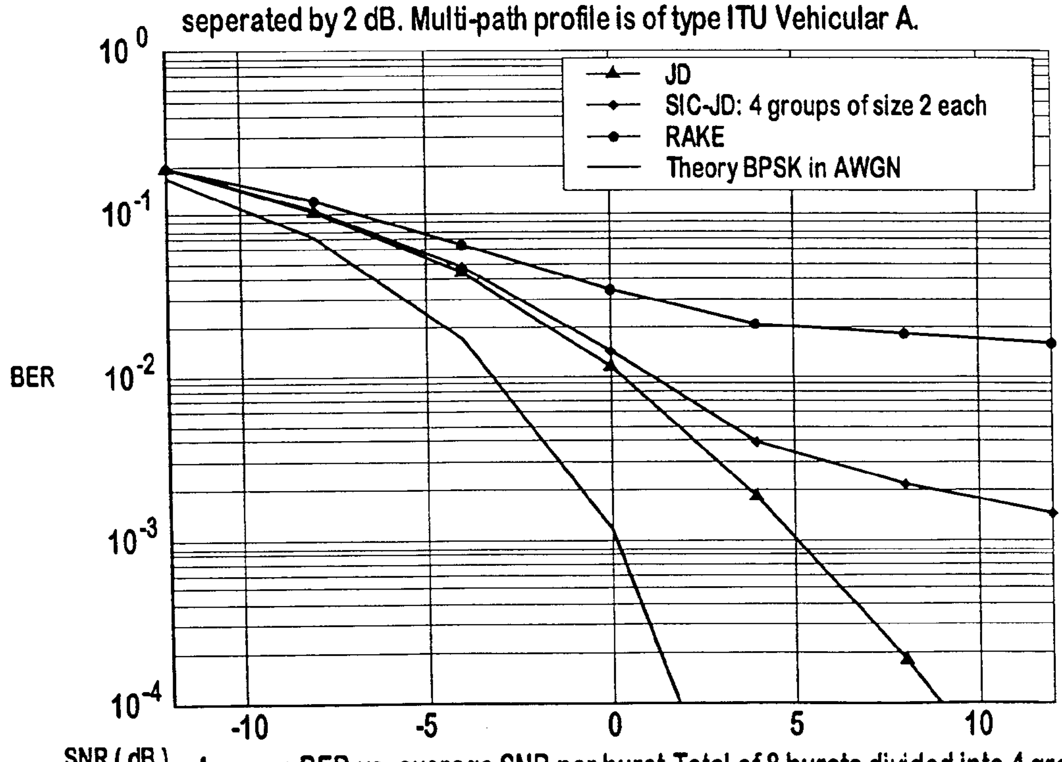


FIG. 11

Average BER vs. average SNR per burst. Total of 8 bursts divided into 4 groups with 2 bursts per group. All bursts in the same group are subjected to the same channel. All 4 groups pass through a common channel but their average SNR is seperated by 2 dB. Multi-path profile is of type 3GPP WG4 Case 2.

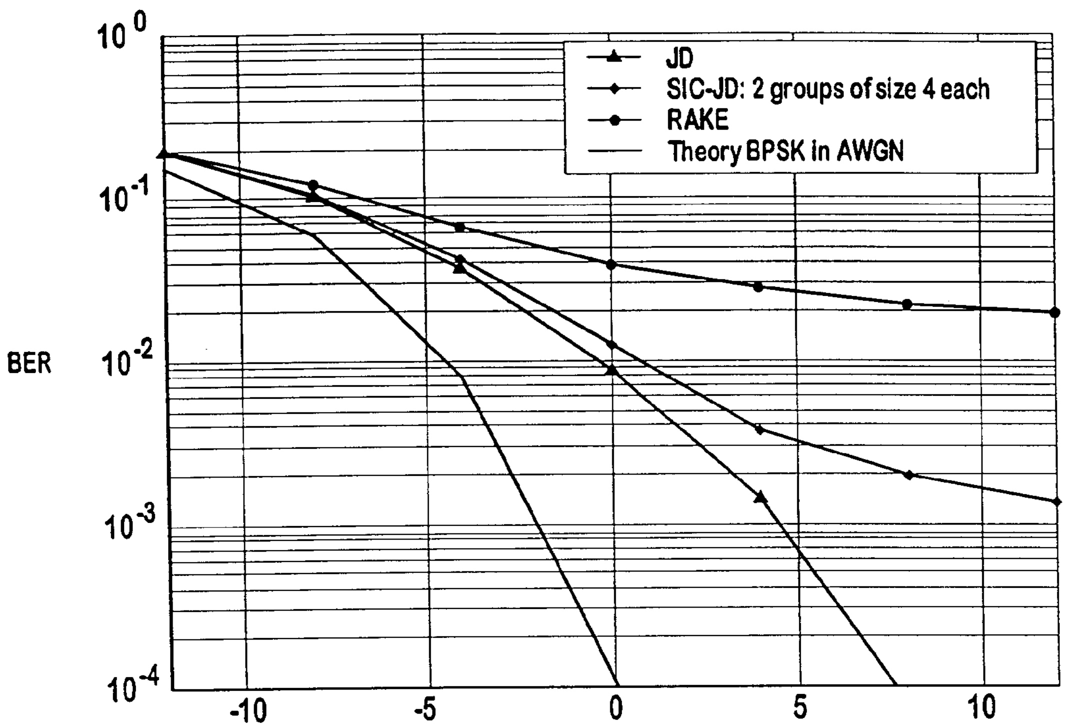


FIG. 12

Average BER vs. average SNR per burst. Total of 8 bursts divided into 2 groups with 4 bursts per group. All bursts in the same group are subjected to the same channel. All groups pass through a common channel but their average SNR is seperated by 2 dB. Multi-path profile is of type ITU Vehicular A.

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
11 October 2001 (11.10.2001)

PCT

(10) International Publication Number
WO 01/76110 A2

(51) International Patent Classification[7]: H04B 17/00

(21) International Application Number: PCT/US01/09114

(22) International Filing Date: 20 March 2001 (20.03.2001)

(25) Filing Language: English

(26) Publication Language: English

(30) Priority Data:
09/539,224 30 March 2000 (30.03.2000) US

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(81) Designated States *(national)*: AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, BZ, CA, CH, CN, CR, CU, CZ, DE, DK, DM, DZ, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NO, NZ, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TR, TT, TZ, UA, UG, UZ, VN, YU, ZA, ZW.

(84) Designated States *(regional)*: ARIPO patent (GH, GM, KE, LS, MW, MZ, SD, SL, SZ, TZ, UG, ZW), Eurasian patent (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European patent (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE, TR), OAPI patent (BF, BJ, CF, CG, CI, CM, GA, GN, GW, ML, MR, NE, SN, TD, TG).

Published:
— *without international search report and to be republished upon receipt of that report*

WO 01/76110 A2

(54) Title: METHOD AND APPARATUS FOR MEASURING AND REPORTING CHANNEL STATE INFORMATION IN A HIGH EFFICIENCY, HIGH PERFORMANCE COMMUNICATIONS SYSTEM

(57) Abstract: Channel state information (CSI) can be used by a communications system to precondition transmissions between transmitter units and receiver units. In one aspect of the invention, disjoint sub-channel sets are assigned to transmit antennas located at a transmitter unit. Pilot symbols are generated and transmitted on a subset of the disjoint sub-channels. Upon receipt of the transmitted pilot symbols, the receiver units determine the CSI for the disjoint sub-channels that carried pilot symbols. These CSI values are reported to the transmitter unit, which will use these CSI values to generate CSI estimates for the disjoint sub-channels that did not carry pilot symbols. The amount of information necessary to report CSI on the reverse link can be further minimized through compression techniques and resource allocation techniques.

METHOD AND APPARATUS FOR MEASURING AND REPORTING CHANNEL STATE INFORMATION IN A HIGH EFFICIENCY, HIGH PERFORMANCE COMMUNICATIONS SYSTEM

BACKGROUND OF THE INVENTION

I. Field of the Invention

The present invention relates to the field of communications. More particularly, the present invention relates to the measurement and report of channel state information in a high efficiency, high performance communications system.

II. Description of the Related Art

A modern day wireless communications system is required to operate over channels that experience fading and multipath. One such communications system is a code division multiple access (CDMA) system that conforms to the "TIA/EIA/IS-95 Mobile Station-Base Station Compatibility Standard for Dual-Mode Wideband Spread Spectrum Cellular System," hereinafter referred to as the IS-95 standard. The CDMA system supports voice and data communication between users over a terrestrial link. The use of CDMA techniques in a multiple access communication system is disclosed in U.S. Patent No. 4,901,307, entitled "SPREAD SPECTRUM MULTIPLE ACCESS COMMUNICATION SYSTEM USING SATELLITE OR TERRESTRIAL REPEATERS," and U.S. Patent No. 5,103,459, entitled "SYSTEM AND METHOD FOR GENERATING WAVEFORMS IN A CDMA CELLULAR TELEPHONE SYSTEM," both assigned to the assignee of the present invention and incorporated herein by reference.

An IS-95 system can operate efficiently by estimating channel parameters at a receiver unit, which uses these estimated channel parameters to demodulate a received signal. The IS-95 system makes channel estimation efficient by requiring the transmission of a pilot signal

from every base station. This pilot signal is a repeating PN-type sequence known by the receiver unit. Correlation of the received pilot signal with a local replica of the pilot signal enables the receiver unit to estimate the complex impulse response of the channel and adjust demodulator parameters accordingly. For the IS-95 waveform and system parameters it is not necessary or beneficial to report information on the channel conditions measured by the receiver unit back to the transmitter unit.

Given the ever-growing demand for wireless communication, a higher efficiency, higher performance wireless communications system is desirable. One type of higher performance wireless communications system is a Multiple Input/Multiple Output (MIMO) system that employs multiple transmit antennas to transmit over a propagation channel to multiple receive antennas. As in lower performance systems, the propagation channel in a MIMO system is subject to the deleterious effects of multipath, as well as interference from adjacent antennas. Multipath occurs when a transmitted signal arrives at a receiver unit through multiple propagation paths with differing delays. When signals arrive from multiple propagation paths, components of the signals can combine destructively, which is referred to as "fading." In order to improve the efficiency and decrease the complexity of the MIMO system, information as to the characteristics of the propagation channel can be transmitted back to the transmitter unit in order to precondition the signal before transmission.

Preconditioning the signal can be difficult when the characteristics of the propagation channel change rapidly. The channel response can change with time due to the movement of the receiver unit or changes in the environment surrounding the receiver unit. Given a mobile environment, an optimal performance requires that information regarding channel characteristics, such as fading and interference statistics, be determined and transmitted quickly to the transmitter unit before the channel characteristics change significantly. As delay of the measurement and reporting process increases, the utility of the channel response information decreases. A

present need exists for efficient techniques that will provide rapid determination of the channel characteristics.

SUMMARY OF THE INVENTION

The present invention is directed to a method and apparatus for the measuring and reporting of channel state information in a high efficiency, high performance communications system, comprising the steps of: generating a plurality of pilot signals; transmitting the plurality of pilot signals over a propagation channel between a transmitter unit and a plurality of receiver units, wherein the transmitter unit comprises at least one transmit antenna, each of the plurality of receiver units comprises at least one receive antenna, and the propagation channel comprises a plurality of sub-channels between the transmitter unit and the plurality of receiver units; receiving at least one of the plurality of pilot signals at each of the plurality of receiver units; determining a set of transmission characteristics for at least one of the plurality of sub-channels, wherein the step of determining the set of transmission characteristics uses at least one of the plurality of pilot signals received at each of the plurality of receiver units; reporting an information signal from each of the plurality of receiver units to the transmitter unit, wherein the information signal carries the set of transmission characteristics for at least one of the plurality of sub-channels; and optimizing a set of transmission parameters at the transmitter unit, based on the information signal.

In one aspect of the invention, pilot symbols are transmitted on a plurality of disjoint OFDM sub-channel sets. When the pilot symbols are transmitted on disjoint OFDM sub-channels, the characteristics of the propagation channel can be determined through a set of K sub-channels carrying the pilot symbols, wherein K is less than the number of OFDM sub-channels in the system. In addition to transmitting pilot symbols on disjoint sub-channels, the system can transmit a time-domain pilot sequence that can be used to determine characteristics of the propagation

channel. Along with the generation and transmission of pilot symbols, an aspect of the invention is the compression of the amount of information necessary to reconstruct the characteristics of the propagation channel.

BRIEF DESCRIPTION OF THE DRAWINGS

The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

FIG. 1A is a diagram of a multiple-input multiple-output (MIMO) communications system;

FIG. 1B is a diagram of a OFDM-based MIMO system with feedback of channel state information;

FIG. 1C is a diagram of an exemplary OFDM pilot signal structure that can be used to estimate the channel state information;

FIG. 2 is a diagram that graphically illustrates a specific example of a transmission from a transmit antenna at a transmitter unit;

FIG. 3 is a block diagram of a data processor and a modulator of the communications system shown in FIG. 1A;

FIGS. 4A and 4B are block diagrams of two versions of a channel data processor that can be used for processing one channel data steam such as control, broadcast, voice, or traffic data;

FIGS. 5A through 5C are block diagrams of the processing units that can be used to generate the transmit signal shown in FIG. 2;

FIG. 6 is a block diagram of a receiver unit, having multiple receive antennas, which can be used to receive one or more channel data streams; and

FIG. 7 shows plots that illustrate the spectral efficiency achievable with some of the operating modes of a communications system in accordance with one embodiment.

DETAILED DESCRIPTION OF THE SPECIFIC EMBODIMENTS

FIG. 1A is a diagram of a Multiple Input/Multiple Output (MIMO) communications system 100 capable of implementing some embodiments of the invention. Communications system 100 can be operative to provide a combination of antenna, frequency, and temporal diversity to increase spectral efficiency, improve performance, and enhance flexibility. Increased spectral efficiency is characterized by the ability to transmit more bits per second per Hertz (bps/Hz) when and where possible to better utilize the available system bandwidth. Techniques to obtain higher spectral efficiency are described in further detail below. Improved performance may be quantified, for example, by a lower bit-error-rate (BER) or frame-error-rate (FER) for a given link carrier-to-noise-plus-interference ratio (C/I). And enhanced flexibility is characterized by the ability to accommodate multiple users having different and typically disparate requirements. These goals may be achieved, in part, by employing multi-carrier modulation, time division multiplexing (TDM), multiple transmit and/or receive antennas, and other techniques. The features, aspects, and advantages of the invention are described in further detail below.

As shown in FIG. 1A, communications system 100 includes a first system 110 in communication with a second system 120. System 110 includes a (transmit) data processor 112 that (1) receives or generates data, (2) processes the data to provide antenna, frequency, or temporal diversity, or a combination thereof, and (3) provides processed modulation symbols to a number of modulators (MOD) 114a through 114t. Each modulator 114 further processes the modulation symbols and generates an RF modulated signal suitable for transmission. The RF modulated signals from modulators 114a through 114t are then transmitted from respective antennas 116a through 116t over communications links 118 to system 120.

In FIG. 1A, system 120 includes a number of receive antennas 122a through 122r that receive the transmitted signals and provide the received signals to respective demodulators (DEMOD) 124a through 124r. As shown in FIG. 1A, each receive antenna 122 may receive signals from one or more

transmit antennas 116 depending on a number of factors such as, for example, the operating mode used at system 110, the directivity of the transmit and receive antennas, the characteristics of the communications links, and others. Each demodulator 124 demodulates the respective received signal using a demodulation scheme that is complementary to the modulation scheme used at the transmitter. The demodulated symbols from demodulators 124a through 124r are then provided to a (receive) data processor 126 that further processes the symbols to provide the output data. The data processing at the transmitter and receiver units is described in further detail below.

FIG. 1A shows only the forward link transmission from system 110 to system 120. This configuration may be used for data broadcast and other one-way data transmission applications. In a bi-directional communications system, a reverse link from system 120 to system 110 is also provided, although not shown in FIG. 1A for simplicity. For the bi-directional communications system, each of systems 110 and 120 may operate as a transmitter unit or a receiver unit, or both concurrently, depending on whether data is being transmitted from, or received at, the unit.

For simplicity, communications system 100 is shown to include one transmitter unit (i.e., system 110) and one receiver unit (i.e., system 120). However, in general, multiple transmit antennas and multiple receive antennas are present on each transmitter unit and each receiver unit. The communications system of the invention may include any number of transmitter units and receiver units.

Each transmitter unit may include a single transmit antenna or a number of transmit antennas, such as that shown in FIG. 1A. Similarly, each receiver unit may include a single receive antenna or a number of receive antennas, again such as that shown in FIG. 1A. For example, the communications system may include a central system (i.e., similar to a base station in the IS-95 CDMA system) having a number of antennas that transmit data to, and receive data from, a number of remote systems (i.e., subscriber units, similar to remote stations in the CDMA system), some of

which may include one antenna and others of which may include multiple antennas.

As used herein, an antenna refers to a collection of one or more antenna elements that are distributed in space. The antenna elements may be physically located at a single site or distributed over multiple sites. Antenna elements physically co-located at a single site may be operated as an antenna array (e.g., such as for a CDMA base station). An antenna network consists of a collection of antenna arrays or elements that are physically separated (e.g., several CDMA base stations). An antenna array or an antenna network may be designed with the ability to form beams and to transmit multiple beams from the antenna array or network. For example, a CDMA base station may be designed with the capability to transmit up to three beams to three different sections of a coverage area (or sectors) from the same antenna array. Thus, the three beams may be viewed as three transmissions from three antennas.

The communications system of the invention can be designed to provide a multi-user, multiple access communications scheme capable of supporting subscriber units having different requirements as well as capabilities. The scheme allows the system's total operating bandwidth, W, (e.g., 1.2288 MHz) to be efficiently shared among different types of services that may have highly disparate data rate, delay, and quality of service (QOS) requirements.

Examples of such disparate types of services include voice services and data services. Voice services are typically characterized by a low data rate (e.g., 8 kbps to 32 kbps), short processing delay (e.g., 3 msec to 100 msec overall one-way delay), and sustained use of a communications channel for an extended period of time. The short delay requirements imposed by voice services typically require a small fraction of the system resources to be dedicated to each voice call for the duration of the call. In contrast, data services are characterized by "bursty" traffics in which variable amounts of data are sent at sporadic times. The amount of data can vary significantly from burst-to-burst and from user-to-user. For high efficiency, the

communications system of the invention can be designed with the capability to allocate a portion of the available resources to voice services as required and the remaining resources to data services. A fraction of the available system resources may also be dedicated for certain data services or certain types of data services.

The distribution of data rates achievable by each subscriber unit can vary widely between some minimum and maximum instantaneous values (e.g., from 200 kbps to over 20 Mbps). The achievable data rate for a particular subscriber unit at any given moment may be influenced by a number of factors such as the amount of available transmit power, the quality of the communications link (i.e., the C/I), the coding scheme, and others. The data rate requirement of each subscriber unit may also vary widely between a minimum value (e.g., 8 kbps, for a voice call) all the way up to the maximum supported instantaneous peak rate (e.g., 20 Mbps for bursty data services).

The percentage of voice and data traffic is typically a random variable that changes over time. In accordance with certain aspects of the invention, to efficiently support both types of services concurrently, the communications system of the invention is designed with the capability to dynamic allocate the available resources based on the amount of voice and data traffic. A scheme to dynamically allocate resources is described below. Another scheme to allocate resources is described in the aforementioned U.S. Patent Application Serial No. 08/963,386.

The communications system of the invention provides the above-described features and advantages, and is capable of supporting different types of services having disparate requirements. The features are achieved by employing antenna, frequency, or temporal diversity, or a combination thereof. Antenna, frequency, or temporal diversity can be independently achieved and dynamically selected.

As used herein, antenna diversity refers to the transmission and/or reception of data over more than one antenna, frequency diversity refers to the transmission of data over more than one sub-band, and temporal

diversity refers to the transmission of data over more than one time period. Antenna, frequency, and temporal diversity may include subcategories. For example, transmit diversity refers to the use of more than one transmit antenna in a manner to improve the reliability of the communications link, receive diversity refers to the use of more than one receive antenna in a manner to improve the reliability of the communications link, and spatial diversity refers to the use of multiple transmit and receive antennas to improve the reliability and/or increase the capacity of the communications link. Transmit and receive diversity can also be used in combination to improve the reliability of the communications link without increasing the link capacity. Various combinations of antenna, frequency, and temporal diversity can thus be achieved and are within the scope of the present invention.

Frequency diversity can be provided by use of a multi-carrier modulation scheme such as orthogonal frequency division multiplexing (OFDM), which allows for transmission of data over various sub-bands of the operating bandwidth. Temporal diversity is achieved by transmitting the data over different times, which can be more easily accomplished with the use of time-division multiplexing (TDM). These various aspects of the communications system of the invention are described in further detail below.

In accordance with an aspect of the invention, antenna diversity is achieved by employing a number of (N_T) transmit antennas at the transmitter unit or a number of (N_R) receive antennas at the receiver unit, or multiple antennas at both the transmitter and receiver units. In a terrestrial communications system (e.g., a cellular system, a broadcast system, an MMDS system, and others), an RF modulated signal from a transmitter unit may reach the receiver unit via a number of transmission paths. The characteristics of the transmission paths typically vary over time based on a number of factors. If more than one transmit or receive antenna is used, and if the transmission paths between the transmit and receive antennas are independent (i.e., uncorrelated), which is generally true to at

least an extent, then the likelihood of correctly receiving the transmitted signal increases as the number of antennas increases. Generally, as the number of transmit and receive antennas increases, diversity increases and performance improves.

Antenna diversity is dynamically provided based on the characteristics of the communications link to provide the required performance. For example, a higher degree of antenna diversity can be provided for some types of communication (e.g., signaling), for some types of services (e.g., voice), for some communications link characteristics (e.g., low C/I), or for some other conditions or considerations.

As used herein, antenna diversity includes transmit diversity and receive diversity. For transmit diversity, data is transmitted over multiple transmit antennas. Typically, additional processing is performed on the data transmitted from the transmit antennas to achieve the desired diversity. For example, the data transmitted from different transmit antennas may be delayed or reordered in time, or coded and interleaved across the available transmit antennas. Also, frequency and temporal diversity may be used in conjunction with the different transmit antennas. For receive diversity, modulated signals are received on multiple receive antennas, and diversity is achieved by simply receiving the signals via different transmission paths.

In accordance with another aspect of the invention, frequency diversity can be achieved by employing a multi-carrier modulation scheme. One such scheme that has numerous advantages is OFDM. With OFDM modulation, the overall transmission channel is essentially divided into a number of (L) parallel sub-channels that are used to transmit the same or different data. The overall transmission channel occupies the total operating bandwidth of W, and each of the sub-channels occupies a sub-band having a bandwidth of W/L and centered at a different center frequency. Each sub-channel has a bandwidth that is a portion of the total operating bandwidth. Each of the sub-channels may also be considered an independent data transmission channel that may be associated with a

particular (and possibly unique) processing, coding, and modulation scheme, as described below.

The data may be partitioned and transmitted over any defined set of two or more sub-bands to provide frequency diversity. For example, the transmission to a particular subscriber unit may occur over sub-channel 1 at time slot 1, sub-channel 5 at time slot 2, sub-channel 2 at time slot 3, and so on. As another example, data for a particular subscriber unit may be transmitted over sub-channels 1 and 2 at time slot 1 (e.g., with the same data being transmitted on both sub-channels), sub-channels 4 and 6 at time slot 2, only sub-channel 2 at time slot 3, and so on. Transmission of data over different sub-channels over time can improve the performance of a communications system experiencing frequency selective fading and channel distortion. Other benefits of OFDM modulation are described below.

In accordance with yet another aspect of the invention, temporal diversity is achieved by transmitting data at different times, which can be more easily accomplished using time division multiplexing (TDM). For data services (and possibly for voice services), data transmission occurs over time slots that may be selected to provide immunity to time dependent degradation in the communications link. Temporal diversity may also be achieved through the use of interleaving.

For example, the transmission to a particular subscriber unit may occur over time slots 1 through x, or on a subset of the possible time slots from 1 through x (e.g., time slots 1, 5, 8, and so on). The amount of data transmitted at each time slot may be variable or fixed. Transmission over multiple time slots improves the likelihood of correct data reception due to, for example, impulse noise and interference.

The combination of antenna, frequency, and temporal diversity allows the communications system of the invention to provide robust performance. Antenna, frequency, and/or temporal diversity improves the likelihood of correct reception of at least some of the transmitted data, which may then be used (e.g., through decoding) to correct for some errors

that may have occurred in the other transmissions. The combination of antenna, frequency, and temporal diversity also allows the communications system to concurrently accommodate different types of services having disparate data rate, processing delay, and quality of service requirements.

The communications system of the invention can be designed and operated in a number of different communications modes, with each communications mode employing antenna, frequency, or temporal diversity, or a combination thereof. The communications modes include, for example, a diversity communications mode and a MIMO communications mode. Various combinations of the diversity and MIMO communications modes can also be supported by the communications system. Also, other communications modes can be implemented and are within the scope of the present invention.

The diversity communications mode employs transmit and/or receive diversity, frequency, or temporal diversity, or a combination thereof, and is generally used to improve the reliability of the communications link. In one implementation of the diversity communications mode, the transmitter unit selects a modulation and coding scheme (i.e., configuration) from a finite set of possible configurations, which are known to the receiver units. For example, each overhead and common channel may be associated with a particular configuration that is known to all receiver units. When using the diversity communications mode for a specific user (e.g., for a voice call or a data transmission), the mode and/or configuration may be known *a priori* (e.g., from a previous set up) or negotiated (e.g., via a common channel) by the receiver unit.

In the diversity communications mode, data is transmitted on one or more sub-channels, from one or more antennas, and at one or more time periods. The allocated sub-channels may be associated with the same antenna, or may be sub-channels associated with different antennas. In a common application of the diversity communications mode, which is also referred to as a "pure" diversity communications mode, data is transmitted from all available transmit antennas to the destination receiver unit. The

pure diversity communications mode can be used in instances where the data rate requirements are low or when the C/I is low, or when both are true.

The MIMO communications mode employs antenna diversity at both ends of the communication link and is generally used to improve both the reliability and increase the capacity of the communications link. The MIMO communications mode may further employ frequency and/or temporal diversity in combination with the antenna diversity. The MIMO communications mode, which may also be referred to herein as the spatial communications mode, employs one or more processing modes to be described below.

The diversity communications mode generally has lower spectral efficiency than the MIMO communications mode, especially at high C/I levels. However, at low to moderate C/I values, the diversity communications mode achieves comparable efficiency and can be simpler to implement. In general, the use of the MIMO communications mode provides greater spectral efficiency when used, particularly at moderate to high C/I values. The MIMO communications mode may thus be advantageously used when the data rate requirements are moderate to high.

The communications system can be designed to concurrently support both diversity and MIMO communications modes. The communications modes can be applied in various manners and, for increased flexibility, may be applied independently on a sub-channel basis. The MIMO communications mode is typically applied to specific users. However, each communications mode may be applied on each sub-channel independently, across a subset of sub-channels, across all sub-channels, or on some other basis. For example, the use of the MIMO communications mode may be applied to a specific user (e.g., a data user) and, concurrently, the use of the diversity communications mode may be applied to another specific user (e.g., a voice user) on a different sub-channel. The diversity communications mode may also be applied, for example, on sub-channels experiencing higher path loss.

The communications system of the invention can also be designed to support a number of processing modes. When the transmitter unit is provided with information indicative of the conditions (i.e., the "state") of the communications links, additional processing can be performed at the transmitter unit to further improve performance and increase efficiency. Full channel state information (CSI) or partial CSI may be available to the transmitter unit. Full CSI includes sufficient characterization of the propagation path (i.e., amplitude and phase) between all pairs of transmit and receive antennas for each sub-band. Full CSI also includes the C/I per sub-band. The full CSI may be embodied in a set of matrices of complex gain values that are descriptive of the conditions of the transmission paths from the transmit antennas to the receive antennas, as described below. Partial CSI may include, for example, the C/I of the sub-band . With full CSI or partial CSI, the transmitter unit pre-conditions the data prior to transmission to receiver unit.

The transmitter unit can precondition the signals presented to the transmit antennas in a way that is unique to a specific receiver unit (e.g., the pre-conditioning is performed for each sub-band assigned to that receiver unit). As long as the channel does not change appreciably from the time it is measured by the receiver unit and subsequently sent back to the transmitter and used to precondition the transmission, the intended receiver unit can demodulate the transmission. In this implementation, a full-CSI based MIMO communication can only be demodulated by the receiver unit associated with the CSI used to precondition the transmitted signals.

In the partial-CSI or no-CSI processing modes, the transmitter unit can employ a common modulation and coding scheme (e.g., on each data channel transmission), which then can be (in theory) demodulated by all receiver units. In the partial-CSI processing mode, a single receiver unit can specify the C/I, and the modulation employed on all antennas can be selected accordingly (e.g., for reliable transmission) for that receiver unit. Other receiver units can attempt to demodulate the transmission and, if they have adequate C/I, may be able to successfully recover the transmission. A common (e.g., broadcast) channel can use a no-CSI processing mode to reach all users.

As an example, assume that the MIMO communications mode is applied to a channel data stream that is transmitted on one particular sub-

channel from four transmit antennas. The channel data stream is demultiplexed into four data sub-streams, one data sub-stream for each transmit antenna. Each data sub-stream is then modulated using a particular modulation scheme (e.g., M-PSK, M-QAM, or other) selected based on the CSI for that sub-band and for that transmit antenna. Four modulation sub-streams are thus generated for the four data sub-streams, with each modulation sub-streams including a stream of modulation symbols. The four modulation sub-streams are then pre-conditioned using the eigenvector matrix, as expressed below in equation (1), to generate pre-conditioned modulation symbols. The four streams of pre-conditioned modulation symbols are respectively provided to the four combiners of the four transmit antennas. Each combiner combines the received pre-conditioned modulation symbols with the modulation symbols for the other sub-channels to generate a modulation symbol vector stream for the associated transmit antenna.

The full-CSI based processing is typically employed in the MIMO communications mode where parallel data streams are transmitted to a specific user on each of the channel eigenmodes for the each of the allocated sub-channels. Similar processing based on full CSI can be performed where transmission on only a subset of the available eigenmodes is accommodated in each of the allocated sub-channels(e.g., to implement beam steering). Because of the cost associated with the full-CSI processing (e.g., increased complexity at the transmitter and receiver units, increased overhead for the transmission of the CSI from the receiver unit to the transmitter unit, and so on), full-CSI processing can be applied in certain instances in the MIMO communications mode where the additional increase in performance and efficiency is justified.

In instances where full CSI is not available, less descriptive information on the transmission path (or partial CSI) may be available and can be used to pre-condition the data prior to transmission. For example, the C/I of each of the sub-channels may be available. The C/I information can then be used to control the transmission from various transmit

antennas to provide the required performance in the sub-channels of interest and increase system capacity.

As used herein, full-CSI based processing modes denote processing modes that use full CSI, and partial-CSI based processing modes denote processing modes that use partial CSI. The full-CSI based processing modes include, for example, the full-CSI MIMO mode that utilizes full-CSI based processing in the MIMO communications mode. The partial-CSI based modes include, for example, the partial-CSI MIMO mode that utilizes partial-CSI based processing in the MIMO communications mode.

In instances where full-CSI or partial-CSI processing is employed to allow the transmitter unit to pre-condition the data using the available channel state information (e.g., the eigenmodes or C/I), feedback information from the receiver unit is required, which uses a portion of the reverse link capacity. Therefore, there is a cost associated with the full-CSI and the partial-CSI based processing modes. The cost should be factored into the choice of which processing mode to employ. The partial-CSI based processing mode requires less overhead and may be more efficient in some instances. The no-CSI based processing mode requires no overhead and may also be more efficient than the full-CSI based processing mode or the partial-CSI based processing mode under some other circumstances.

FIG. 2 is a diagram that graphically illustrates at least some of the aspects of the communications system of the invention. FIG. 2 shows a specific example of a transmission from one of N_T transmit antennas at a transmitter unit. In FIG. 2, the horizontal axis is time and the vertical axis is frequency. In this example, the transmission channel includes 16 sub-channels and is used to transmit a sequence of OFDM symbols, with each OFDM symbol covering all 16 sub-channels (one OFDM symbol is indicated at the top of FIG. 2 and includes all 16 sub-bands). A TDM structure is also illustrated in which the data transmission is partitioned into time slots, with each time slot having the duration of, for example, the length of one modulation symbol (i.e., each modulation symbol is used as the TDM interval).

The available sub-channels can be used to transmit signaling, voice, traffic data, and others. In the example shown in FIG. 2, the modulation symbol at time slot 1 corresponds to pilot data, which is periodically transmitted to assist the receiver units to synchronize and perform channel estimation. Other techniques for distributing pilot data over time and frequency can also be used and are within the scope of the present invention. In addition, it may be advantageous to utilize a particular modulation scheme during the pilot interval if all sub-channels are employed (e.g., a PN code with a chip duration of approximately 1/W). Transmission of the pilot modulation symbol typically occurs at a particular frame rate, which is usually selected to be fast enough to permit accurate tracking of variations in the communications link.

The time slots not used for pilot transmissions can then be used to transmit various types of data. For example, sub-channels 1 and 2 may be reserved for the transmission of control and broadcast data to the receiver units. The data on these sub-channels is generally intended to be received by all receiver units. However, some of the messages on the control channel may be user specific, and can be encoded accordingly.

Voice data and traffic data can be transmitted in the remaining sub-channels. For the example shown in FIG. 2, sub-channel 3 at time slots 2 through 9 is used for voice call 1, sub-channel 4 at time slots 2 through 9 is used for voice call 2, sub-channel 5 at time slots 5 through 9 is used for voice call 3, and sub-channel 6 at time slots 7 through 9 is used for voice call 5.

The remaining available sub-channels and time slots may be used for transmissions of traffic data. In the example shown in FIG. 2, data 1 transmission uses sub-channels 5 through 16 at time slot 2 and sub-channels 7 through 16 at time slot 7, data 2 transmission uses sub-channels 5 through 16 at time slots 3 and 4 and sub-channels 6 through 16 at time slots 5, data 3 transmission uses sub-channels 6 through 16 at time slot 6, data 4 transmission uses sub-channels 7 through 16 at time slot 8, data 5 transmission uses sub-channels 7 through 11 at time slot 9, and data 6 transmission uses sub-channels 12 through 16 at time slot 9. Data 1 through

6 transmissions can represent transmissions of traffic data to one or more receiver units.

The communications system of the invention flexibly supports the transmissions of traffic data. As shown in FIG. 2, a particular data transmission (e.g., data 2) may occur over multiple sub-channels and/or multiple time slots, and multiple data transmissions (e.g., data 5 and 6) may occur at one time slot. A data transmission (e.g., data 1) may also occur over non-contiguous time slots. The system can also be designed to support multiple data transmissions on one sub-channel. For example, voice data may be multiplexed with traffic data and transmitted on a single sub-channel.

The multiplexing of the data transmissions can potentially change from OFDM symbol to symbol. Moreover, the communications mode may be different from user to user (e.g., from one voice or data transmission to other). For example, the voice users may use the diversity communications mode, and the data users may use the MIMO communications modes. This concept can be extended to the sub-channel level. For example, a data user may use the MIMO communications mode in sub-channels that have sufficient C/I and the diversity communications mode in remaining sub-channels.

Antenna, frequency, and temporal diversity may be respectively achieved by transmitting data from multiple antennas, on multiple sub-channels in different sub-bands, and over multiple time slots. For example, antenna diversity for a particular transmission (e.g., voice call 1) may be achieved by transmitting the (voice) data on a particular sub-channel (e.g., sub-channel 1) over two or more antennas. Frequency diversity for a particular transmission (e.g., voice call 1) may be achieved by transmitting the data on two or more sub-channels in different sub-bands (e.g., sub-channels 1 and 2). A combination of antenna and frequency diversity may be obtained by transmitting data from two or more antennas and on two or more sub-channels. Temporal diversity may be achieved by transmitting data over multiple time slots. For example, as shown in FIG. 2, data 1

transmission at time slot 7 is a portion (e.g., new or repeated) of the data 1 transmission at time slot 2.

The same or different data may be transmitted from multiple antennas and/or on multiple sub-bands to obtain the desired diversity. For example, the data may be transmitted on: (1) one sub-channel from one antenna, (2) one sub-channel (e.g., sub-channel 1) from multiple antennas, (3) one sub-channel from all N_T antennas, (4) a set of sub-channels (e.g., sub-channels 1 and 2) from one antenna, (5), a set of sub-channels from multiple antennas, (6) a set of sub-channels from all N_T antennas, or (7) a set of channels from a set of antennas (e.g., sub-channel 1 from antennas 1 and 2 at one time slot, sub-channels 1 and 2 from antenna 2 at another time slot, and so on). Thus, any combination of sub-channels and antennas may be used to provide antenna and frequency diversity.

In accordance with certain embodiments of the invention that provide the most flexibility and are capable of achieving high performance and efficiency, each sub-channel at each time slot for each transmit antenna may be viewed as an independent unit of transmission (i.e., a modulation symbol) that can be used to transmit any type of data such as pilot, signaling, broadcast, voice, traffic data, and others, or a combination thereof (e.g., multiplexed voice and traffic data). In such design, a voice call may be dynamically assigned different sub-channels over time.

Flexibility, performance, and efficiency are further achieved by allowing for independence among the modulation symbols, as described below. For example, each modulation symbol may be generated from a modulation scheme (e.g., M-PSK, M-QAM, and others) that results in the best use of the resource at that particular time, frequency, and space.

A number of constraints may be placed to simplify the design and implementation of the transmitter and receiver units. For example, a voice call may be assigned to a particular sub-channel for the duration of the call, or until such time as a sub-channel reassignment is performed. Also, signaling and/or broadcast data may be designated to some fixed sub-channels (e.g., sub-channel 1 for control data and sub-channel 2 for broadcast

data, as shown in FIG. 2) so that the receiver units know a priori which sub-channels to demodulate to receive the data.

Also, each data transmission channel or sub-channel may be restricted to a particular modulation scheme (e.g., M-PSK, M-QAM) for the duration of the transmission or until such time as a new modulation scheme is assigned. For example, in FIG. 2, voice call 1 on sub-channel 3 may use QPSK, voice call 2 on sub-channel 4 may use 16-QAM, data 1 transmission at time slot 2 may use 8-PSK, data 2 transmission at time slots 3 through 5 may use 16-QAM, and so on.

The use of TDM allows for greater flexibility in the transmission of voice data and traffic data, and various assignments of resources can be contemplated. For example, a user can be assigned one sub-channel for each time slot or, equivalently, four sub-channels every fourth time slot, or some other allocations. TDM allows for data to be aggregated and transmitted at designated time slot(s) for improved efficiency.

If voice activity is implemented at the transmitter, then in the intervals where no voice is being transmitted, the transmitter may assign other users to the sub-channel so that the sub-channel efficiency is maximized. In the event that no data is available to transmit during the idle voice periods, the transmitter can decrease (or turn-off) the power transmitted in the sub-channel, reducing the interference levels presented to other users in the system that are using the same sub-channel in another cell in the network. The same feature can be also extended to the overhead, control, data, and other channels.

Allocation of a small portion of the available resources over a continuous time period typically results in lower delays, and may be better suited for delay sensitive services such as voice. Transmission using TDM can provide higher efficiency, at the cost of possible additional delays. The communications system of the invention can allocate resources to satisfy user requirements and achieve high efficiency and performance.

Measuring and Reporting Channel State Information In A MIMO System

Given the complexity of a system using multiple transmit antennas and multiple receive antennas, with the associated dispersive channel effects, the preferred modulation technique is OFDM, which effectively decomposes the channel into a set of non-interfering narrowband channels, or sub-channels. With proper OFDM signal design, a signal transmitted on one subchannel sees "flat fading", i.e., the channel response is effectively constant over the subchannel bandwidth. The channel state information or CSI includes sufficient characterization of the propagation path (i.e., amplitude and phase) between all pairs of transmit and receive antennas for each sub-channel. CSI also includes the information of the relative levels of interference and noise in each sub-channel, that is known as C/I information. The CSI may be embodied in a set of matrices of complex gain values that are descriptive of the conditions of the transmission paths from the transmit antennas to the receive antennas, as described below. With CSI, the transmitter unit pre-conditions the data prior to transmission to receiver unit.

CSI processing is briefly described below. When the CSI is available at the transmitter unit, a simple approach is to decompose the multi-input/multi-output channel into a set of independent channels. Given the channel transfer function at the transmitters, the left eigenvectors may be used to transmit different data streams. The modulation alphabet used with each eigenvector is determined by the available C/I of that mode, given by the eigenvalues. If $\mathbf{H}$ is the N_R x N_T matrix that gives the channel response for the N_T transmitter antenna elements and N_R receiver antenna elements at a specific time, and $\underline{\mathbf{x}}$ is the N_T-vector of inputs to the channel, then the received signal can be expressed as:

$$\underline{\mathbf{y}} = \underline{\mathbf{H}}\underline{\mathbf{x}} + \underline{\mathbf{n}},$$

where $\underline{\mathbf{n}}$ is an N_R-vector representing noise plus interference. The eigenvector decomposition of the Hermitian matrix formed by the product of the channel matrix with its conjugate-transpose can be expressed as:

$$\mathbf{H}^*\mathbf{H} = E\ddot{\mathbf{E}}E^* \ ,$$

where the symbol * denotes conjugate-transpose, E is the eigenvector matrix, and $\ddot{\mathbf{E}}$ is a diagonal matrix of eigenvalues, both of dimension N_TxN_T.The transmitter converts a set of N_T modulation symbols $\underline{\mathbf{b}}$ using the eigenvector matrix E. The transmitted modulation symbols from the N_T transmit antennas can thus be expressed as:

$$\underline{\mathbf{x}} = E\underline{\mathbf{b}} \ .$$

For all antennas, the pre-conditioning can thus be achieved by a matrix multiply operation expressed as:

$$\begin{bmatrix} x_1 \\ x_2 \\ \vdots \\ x_{N_T} \end{bmatrix} = \begin{bmatrix} e_{11}, & e_{12}, & & e_{1N_T} \\ e_{21}, & e_{22}, & & e_{2N_T} \\ & & & \\ e_{N_T1}, & e_{N_T1}, & & e_{N_TN_T} \end{bmatrix} \bullet \begin{bmatrix} b_1 \\ b_2 \\ \vdots \\ b_{N_T} \end{bmatrix} \qquad \text{Eq (2)}$$

where b_1, b_2, ... and b_{NT} are respectively the modulation symbols for a particular sub-channel at transmit antennas 1, 2, ...N_T, where each modulation symbol can be generated using, for example, M-PSK, M-QAM, and so on, as described below;

E = is the eigenvector matrix related to the transmission loss from transmit antennas to the receive antennas; and

x_1, x_2, ... x_{NT} are the pre-conditioned modulation symbols, which can be expressed as:

$$x_1 = b_1 \bullet e_{11} + b_2 \bullet e_{12} + \ \ldots \ + b_{N_T} \bullet e_{1N_T} \ ,$$

$$x_2 = b_1 \bullet e_{21} + b_2 \bullet e_{22} + \ \ldots \ + b_{N_T} \bullet e_{2N_T} \ , \text{and}$$

$$x_{N_T} = b_1 \bullet e_{N_T 1} + b_2 \bullet e_{N_T 2} + \ \ldots \ + b_{N_T} \bullet e_{N_T N_T} \quad .$$

Since **H*H** is Hermitian, the eigenvector matrix is unitary. Thus, if the elements of $\underline{\mathbf{b}}$ have equal power, the elements of $\underline{\mathbf{x}}$ also have equal power. The received signal may then be expressed as:

$$\underline{\mathbf{y}} = \mathbf{HE}\underline{\mathbf{b}} + \underline{\mathbf{n}} \quad .$$

The receiver performs a channel-matched-filter operation, followed by multiplication by the right eigenvectors. The result of the channel-matched-filter operation is the vector $\underline{\mathbf{z}}$, which can be expressed as:

$$\underline{\mathbf{z}} = E^*\mathbf{H}^*\mathbf{H}E\underline{\mathbf{b}} + E^*\mathbf{H}^*\underline{\mathbf{n}} = \ddot{\mathbf{E}}\underline{\mathbf{b}} + \underline{\hat{\mathbf{n}}} \ , \qquad \text{Eq.(2)}$$

where the new noise term has covariance that can be expressed as:

$$E(\underline{\hat{n}\hat{n}^*}) = E(E^*\mathbf{H}^*\underline{\mathbf{nn}}^*\mathbf{H}E) = E^*\mathbf{H}^*\mathbf{H}E = \Lambda \ ,$$

i.e., the noise components are independent with variance given by the eigenvalues. The C/I of the i-th component of $\underline{\mathbf{z}}$ is λ_i , the i-th diagonal element of $\ddot{\mathbf{E}}$.

The transmitter unit can thus select a modulation alphabet (i.e., signal constellation) for each of the eigenvectors based on the C/I that is given by the eigenvalue. Provided that the channel conditions do not change appreciably in the interval between the time the CSI is measured at the receiver and reported and used to precondition the transmission at the transmitter, the performance of the communications system will be equivalent to that of a set of independent AWGN channels with known C/I's.

Such a system is illustrated in FIG. 1B. At step 141, the transmitter unit 140 converts data into multiple data sub-channels. Different QAM constellations are employed, depending upon the SNR of the mode and sub-channel. The data for each sub-channel is preconditioned by the eigenmode

matrix for that sub-channel. At step 142, the preconditioned data for a particular antenna undergoes an inverse-Fast Fourier Transform (IFFT) operation to produce a time-domain signal. At step 143, a cyclic extension or a cyclic prefix is appended to the time-domain signal in order to maintain orthogonality among the OFDM sub-channels in the presence of time-dispersion in the propagation channel. One extended symbol value is generated for each OFDM sub-channel and will be referred to hereafter as an OFDM symbol. At step 144, the OFDM symbols are transmitted from the multiple transmit antennas.

Multiple antennas at a receiver unit 145 receive signals at step 146. At step 147, the received signals undergo a Discrete Fourier Transform (DFT) operation to channelize the received signals. At step 148, the data from each subchannel over all of the receive antennas is processed. At this processing step, information regarding channel characteristics is extracted from the data, and converted into a more compressed format. One compression technique is the use of the conjugate channel response and the eigenmode matrix to reduce the amount of information needed to describe channel characteristics. At step 149, a message containing the compressed channel state information is transmitted from the receiver unit 145 to the transmitter unit 140, which will then be used to precondition further transmissions.

To facilitate the derivation of the CSI, the transmit waveform is made up of known pilot symbols for an initial preamble. The pilot waveforms for different transmit antennas comprise disjoint sets of OFDM subchannels as illustrated for the case when $N_t = 4$ in FIG. 1C.

With OFDM modulation, the propagation channel is divided into L parallel sub-channels. In order to determine the CSI quickly, an initial preamble consisting entirely of known symbols is transmitted. In order to efficiently distinguish the differing channel responses of the different transmit-receive antenna patterns, the pilot signals are assigned disjoint subsets of sub-channels. FIG. 1C is a diagram of an exemplary OFDM pilot structure composed of disjoint sub-channel subsets. A sub-channel set

composed of sub-channels {0, 1, 2, . . ., 2^n-1} is decomposed into four disjoint sub-channel subsets A = {0, 4, 8, . . ., 2^n-4}, B = {1, 5, 9, . . . , 2^n-3}, C = {2, 6, 10, . . ., 2^n-2} and D = {3, 7, 11, . . ., 2^n-1}. Sub-channel subset A 150 is transmitted on transmit antenna Tx1 151, sub-channel subset B 152 is transmitted on transmit antenna Tx2 153, sub-channel subset C 154 is transmitted on transmit antenna Tx3 155, and sub-channel subset D 156 is transmitted on transmit antenna Tx4 157. Generally, each transmit antenna transmits on every N^{th} sub-channel across the channel so that all sub-channels are disjoint between transmit antennas. Known pilot symbols can be transmitted on all sub-channels in a sub-channel subset. The minimum spacing between the sub-channels used by a particular transmit antenna is a function of the channel parameters. If the channel response has a large delay spread, then a close spacing may be necessary. If the number of antennas is large enough that the required spacing may not be achieved for all users with a single OFDM symbol, then a number of consecutive OFDM symbols may be employed, with each antenna assigned a disjoint subset of sub-channels on one or more of the multiple pilot symbols.

From each transmit antenna at a transmitter unit, the receiver unit receives pilot symbols on disjoint sub-channels and makes determinations as to channel characteristics of the disjoint sub-channels. As discussed previously, the receiver unit may have one or more receive antennas. Suppose $\underline{x} = \{x_i, i = 1, . . ., K\}$ are the pilot symbol values that are to be transmitted on K pilot sub-channels for a single transmit antenna. The receiver unit will receive the values $y_{ij} = h_{ij}x_i + n_{ij}$, wherein h_{ij} is the complex channel response for the i^{th} pilot sub-channel received at the j^{th} receive antenna, and n_{ij} is noise. From this relationship, the receiver unit can determine noisy estimates of the channel response of K sub-channels of a single transmit antenna. These noisy estimates may be used to derive estimates for all sub-channels of the propagation channel through a number of different methods, such as simple interpolation to more complex estimation using *a priori* information on the channel dispersion and noise level. The estimates may be improved by transmitting pilot symbols over

consecutive OFDM symbols and then averaging the estimates for each consecutive OFDM symbol.

Estimates are generated at each receive antenna for each transmit antenna broadcasting pilot symbols. The CSI for the complete propagation channel can be represented by the set of channel response matrices $\{H_i, i = 1, 2, \ldots, 2^n\}$, where matrix H_i is associated with the i^{th} sub-channel, and the elements of each matrix H_i are $\{h_{ijk}, j = 1,\ldots,N_r, k = 1,\ldots,N_t\}$, the complex channel response values for each of the N_t transmit and N_r receive antennas.

The use of disjoint sub-channel subsets can further be applied in a system wherein multiple links, e.g., a propagation channel from a transmitter unit to one or more receiver units, are located in close proximity. In a system where a base station transmits signals according to sectors, the transmission area of a sector can overlap the transmission area of another sector. In an ideal base station, transmit antennas in each sector transmit signals in a direction that is completely disjoint from the directions assigned to the transmit antennas of the other sectors. Unfortunately, overlapping areas exist in most sectored base stations. Using this embodiment of the invention, all transmit antennas of a base station are assigned disjoint subsets of sub-channels to avoid interference between the sectors of that base station. Similarly, neighboring base stations may also be the cause of significant interference, and disjoint sets of sub-channels may be assigned among base stations.

In general, the computation of the channel response can be made for every link that is assigned a disjoint sub-channel subset, in the same manner as the response is computed for the principle link. However, a reduced amount of CSI from these interfering links may be reported to the transmitter unit. For example, information as to the average total interference level of neighboring links can be transmitted and used to determine the supportable data rate of the principle link. If several interfering links dominate the average total interference level, then the interference information of these links may be reported individually to the

system in order to determine a more efficient grouping of sub-channels in each disjoint sub-channel subset.

Other CSI information that can be conveyed to the transmitter unit is the total measured power in sub-channels not assigned to the principal link. The total measured power of sub-channels assigned to neighboring links gives an estimate of the total interference plus noise power. If several OFDM symbols are used as the pilot symbol, then the mean measured channel response and the actual received signal values may be used to make a direct estimate of the total noise in a given sub-channel.

In general, the assignment of sub-channels for a network of base stations should follow a pattern of "frequency-reuse," wherein the same sub-channels are used only when the links are sufficiently separated by distance. If a large number of links are interfering with each other, then the number of OFDM sub-channels may be inadequate to allow the assignment of sub-channels for every pilot OFDM symbol. In this circumstance, transmit antennas may be assigned sub-channels for every P-th pilot symbol, where P is an integer value greater than one (1).

In another embodiment of the invention, the OFDM scheme is designed to create OFDM symbol values that minimize or eliminate interference between transmit antennas that use either identical sub-channels or disjoint sub-channels. An orthogonal code, such as Walsh coding, can be used to transform Q pilot signals into Q orthogonal signals representative of the pilot signals. In the case where a Walsh code is used, the number of pilot signals would be a power of two. The use of orthogonal codes can be used together with the previously discussed disjoint sub-channel subsets in order to reduce interference from neighboring links. For example, in a 4×4 MIMO system with a system bandwidth of approximately 1MHz, assume that 256 OFDM sub-channels are to be used. If the multipath is limited to ten microseconds, the disjoint sub-channels carrying pilot symbols should be spaced approximately 50kHz apart or closer. Each sub-channel is approximately 4kHz wide so that a spacing of twelve sub-channels is 48kHz wide. If the OFDM sub-channels are divided into twelve

sets of twenty sub-channels each, sixteen are left unused. Two consecutive OFDM symbols are used as a pilot signal, and orthogonal coding on these two symbols is employed. Hence, there are twenty-four different orthogonal pilot assignments. These twenty-four orthogonal pilots are assigned to different transmit antennas and links to minimize interference.

In another embodiment of the invention, a large number of periodic OFDM symbols can be used as pilot data. The number of OFDM symbols must be large enough so that accurate measurements of interference levels from a large number of different transmit antennas can be made. These average interference levels would be used to set up system-wide constraints on simultaneous transmissions from various sites, i.e., an adaptive blanking scheme to allow all users nearly equivalent performance.

In an alternate embodiment of the invention, the CSI of a MIMO propagation channel can be determined and transmitted for a MIMO system that does not utilize OFDM symbols as pilot signals. Instead, a Maximal-Length Shift Register sequence (m-sequence) can be used to sound the propagation channel. An m-sequence is the output of a shift register with feedback. M-sequences have desirable autocorrelation properties, including the property that correlation over a full period of the sequence with any non-zero circular shift of the sequence yields the value –1, wherein the sequence values are +/-1. Hence, the correlation at zero shift is R, wherein R is the length of the sequence. In order to maintain desirable properties such as correlation in the presence of multipath, a portion of the sequence equal to the delay spread of the channel must be repeated.

For example, if it is known that the channel multipath is limited to some time τ_m and the length of the pilot sequence is at least $R\tau_m$, then R different shifts of the same m-sequence may be used with only minimal mutual interference. These R different shifts are assigned to different transmit antennas of a base station and other base stations that are the cause of major interference.

Links in the MIMO system that are distantly separated can be assigned different m-sequences. The cross-correlation properties of different m-

sequences do not exhibit the minimal correlation properties of a single sequence and its shifts, but different m-sequences behave more or less like random sequences and provide an average correlation level of $\sqrt{R}$ where R is the sequence length. This average correlation level is generally adequate for use in a MIMO system, because of the separation between the links.

A shift register with feedback generates all possible m-sequences, so that sequences are merely shifted versions of a single code word of length $R = 2^m-1$, where m is a positive integer value. Hence, a limited number of different binary m-sequences exist. In order to avoid reuse of the same m-sequence in an area where significant interference may result, filtered versions of longer m-sequences can be used. A filtered version of an m-sequence is no longer binary, but will still display the same basic correlation properties.

For example, suppose that the pilot sequence is to be transmitted at a 1MHz rate, and that the multipath is limited to ten microseconds. Assume that a base station has three sectors, wherein four transmit antennas are assigned to each sector for a total of twelve transmit antennas per site. If a length 127 m-sequence is employed, then twelve different shifts of the sequence may be assigned to the antennas of a single base station, with relative shifts of ten samples each. The total length of the transmitted pilot is then 137 microseconds, which is a full period of the sequence plus ten additional samples to accommodate the multipath spread. Then different base stations can be assigned different m-sequences, with m-sequences repeated in a code reuse pattern designed to minimize the effects of interference from the same m-sequence.

The embodiments of the invention discussed herein have been directed to the design and transmission of pilot signals that will allow a person skilled in the art to derive characteristics of the propagation channel and to report such characteristics to the transmission site. However, the full CSI is a large amount of information and also highly redundant. Many methods are available for compressing the amount of CSI information to be transmitted. One method discussed previously is the use of the Hermatian

matrix H*H, wherein H is the channel response as determined at the receiver unit. The Hermation matrix H*H can be reported to the transmitter unit and be used to precondition transmissions. Due to the properties of Hermitian matrices, only half of the matrix elements need to be transmitted, such as the complex lower triangular portion of the matrix H*H, and the real-valued diagonal. Additional efficiencies are realized if the number of receive antennas is larger than the number of transmit antennas. Another method to reduce the amount of information transmitted to the transmitter unit on the reverse link is to report only a subset of the channel response matrices H_i to the transmitter unit, from which the unreported channel response matrices can be determined through interpolation schemes. In another method, a functional representation of the channel response across the sub-channels may be derived for each transmit/receive antenna pair, e.g., a polynomial function representative of the channel response can be generated. The coefficients of the polynomial function are then transmitted to the transmitter unit.

As an alternative to these methods for compressing CSI information, one embodiment of the invention is directed to the transmission of a time-domain representation of the channel response, which is the channel impulse response. If a time-domain representation of the channel response is simple, as in cases where there are only two or three multipath components, an inverse FFT can be performed upon the set of channel frequency responses. The inverse FFT operation can be performed for each link between a transmit/receive antenna pair. The resulting channel impulse responses are then translated into a set of amplitudes and delays that are reported to the transmitter.

As discussed previously, there is a cost associated with the transmission of CSI in the reverse link, which is reduced when the above embodiments of the invention are implemented in the MIMO system. Another method for reducing the cost is to select users according to the short term average of their CSI requirements. The CSI requirements change as the channel fades, so improved efficiency on the reverse link is achieved

if users estimate the quantity of CSI required, and inform the base station at intervals that may be periodic or aperiodic, depending on the rate of change of the propagation channel observed by the user. The base station may then include this factor in scheduling the use of the forward and reverse links. Scheduling can be arranged so that users associated with slowly changing propagation channels report less frequently than users associated with quickly changing propagation channels. The base station can also arrange the scheduling to take into account factors such as the number of system users and fairness.

In another aspect of this embodiment of the invention, a time interval can be assigned so that CSI updates in a long transmission period can be adjusted according to the actual changes in the propagation channel. Changes in the propagation channel can be monitored at the receiving site in one of a number of possible ways. For example, the difference between the soft decision on the symbols and the closest QAM constellation value can be determined and used as a criterion, or the relative sizes of decoder metrics can also be used. When the quality of a given criterion falls below a predetermined threshold, an update to the CSI is reported to the transmitter unit.

The overall multipath power-delay profile of a link changes very slowly because the average power observed at various delays remains constant, even though channel fading may occur frequently. Hence, the amount of CSI required to characterize a link can vary substantially from link to link. To optimize performance, the coding of the CSI is tailored to the specific link requirements. If the CSI is sent in frequency-domain form, i.e., a set of channel response matrices which are to be interpolated, then links with little multipath require only a small set of channel response matrices.

Structural Components of a High Efficiency, High Performance Communication System

FIG. 3 is a block diagram of a data processor 112 and modulator 114 of system 110 in FIG. 1A. The aggregate input data stream that includes all data to be transmitted by system 110 is provided to a demultiplexer (DEMUX) 310 within data processor 112. Demultiplexer 310 demultiplexes the input data stream into a number of (K) channel data streams, S_1 through S_k. Each channel data stream may correspond to, for example, a signaling channel, a broadcast channel, a voice call, or a traffic data transmission. Each channel data stream is provided to a respective encoder 312 that encodes the data using a particular encoding scheme.

The encoding may include error correction coding or error detection coding, or both, used to increase the reliability of the link. More specifically, such encoding may include, for example, interleaving, convolutional coding, Turbo coding, Trellis coding, block coding (e.g., Reed-Solomon coding), cyclic redundancy check (CRC) coding, and others. Turbo encoding is described in further detail in U.S. Patent Application Serial No. 09/205,511, filed December 4, 1998 entitled "Turbo Code Interleaver Using Linear Congruential Sequences" and in a document entitled "The cdma2000 ITU-R RTT Candidate Submission," hereinafter referred to as the IS-2000 standard, both of which are incorporated herein by reference.

The encoding can be performed on a per channel basis, i.e., on each channel data stream, as shown in FIG. 3. However, the encoding may also be performed on the aggregate input data stream, on a number of channel data streams, on a portion of a channel data stream, across a set of antennas, across a set of sub-channels, across a set of sub-channels and antennas, across each sub-channel, on each modulation symbol, or on some other unit of time, space, and frequency. The encoded data from encoders 312a through 312k is then provided to a data processor 320 that processes the data to generate modulation symbols.

In one implementation, data processor 320 assigns each channel data stream to one or more sub-channels, at one or more time slots, and on one or more antennas. For example, for a channel data stream corresponding to

a voice call, data processor 320 may assign one sub-channel on one antenna (if transmit diversity is not used) or multiple antennas (if transmit diversity is used) for as many time slots as needed for that call. For a channel data stream corresponding to a signaling or broadcast channel, data processor 320 may assign the designated sub-channel(s) on one or more antennas, again depending on whether transmit diversity is used. Data processor 320 then assigns the remaining available resources for channel data streams corresponding to data transmissions. Because of the bursty nature of data transmissions and the greater tolerance to delays, data processor 320 can assign the available resources such that the system goals of high performance and high efficiency are achieved. The data transmissions are thus "scheduled" to achieve the system goals.

After assigning each channel data stream to its respective time slot(s), sub-channel(s), and antenna(s), the data in the channel data stream is modulated using multi-carrier modulation. OFDM modulation is used to provide numerous advantages. In one implementation of OFDM modulation, the data in each channel data stream is grouped to blocks, with each block having a particular number of data bits. The data bits in each block are then assigned to one or more sub-channels associated with that channel data stream.

The bits in each block are then demultiplexed into separate sub-channels, with each of the sub-channels conveying a potentially different number of bits (i.e., based on C/I of the sub-channel and whether MIMO processing is employed). For each of these sub-channels, the bits are grouped into modulation symbols using a particular modulation scheme (e.g., M-PSK or M-QAM) associated with that sub-channel. For example, with 16-QAM, the signal constellation is composed of 16 points in a complex plane (i.e., a + j*b), with each point in the complex plane conveying 4 bits of information. In the MIMO processing mode, each modulation symbol in the sub-channel represents a linear combination of modulation symbols, each of which may be selected from a different constellation.

The collection of L modulation symbols form a modulation symbol vector V of dimensionality L. Each element of the modulation symbol vector V is associated with a specific sub-channel having a unique frequency or tone on which the modulation symbols is conveyed. The collection of these L modulation symbols are all orthogonal to one another. At each time slot and for each antenna, the L modulation symbols corresponding to the L sub-channels are combined into an OFDM symbol using an inverse fast Fourier transform (IFFT). Each OFDM symbol includes data from the channel data streams assigned to the L sub-channels.

OFDM modulation is described in further detail in a paper entitled "Multicarrier Modulation for Data Transmission : An Idea Whose Time Has Come," by John A.C. Bingham, IEEE Communications Magazine, May 1990, which is incorporated herein by reference.

Data processor 320 thus receives and processes the encoded data corresponding to K channel data streams to provide N_T modulation symbol vectors, V_1 through V_{NT}, one modulation symbol vector for each transmit antenna. In some implementations, some of the modulation symbol vectors may have duplicate information on specific sub-channels intended for different transmit antennas. The modulation symbol vectors V_1 through V_{NT} are provided to modulators 114a through 114t, respectively.

In FIG. 3, each modulator 114 includes an IFFT 330, cycle prefix generator 332, and an upconverter 334. IFFT 330 converts the received modulation symbol vectors into their time-domain representations called OFDM symbols. IFFT 330 can be designed to perform the IFFT on any number of sub-channels (e.g., 8, 16, 32, and so on). Alternatively, for each modulation symbol vector converted to an OFDM symbol, cycle prefix generator 332 repeats a portion of the time-domain representation of the OFDM symbol to form the transmission symbol for the specific antenna. The cyclic prefix insures that the transmission symbol retains its orthogonal properties in the presence of multipath delay spread, thereby improving performance against deleterious path effects, as described below. The

implementation of IFFT 330 and cycle prefix generator 332 is known in the art and not described in detail herein.

The time-domain representations from each cycle prefix generator 332 (i.e., the transmission symbols for each antenna) are then processed by upconverter 332, converted into an analog signal, modulated to a RF frequency, and conditioned (e.g., amplified and filtered) to generate an RF modulated signal that is then transmitted from the respective antenna 116.

FIG. 3 also shows a block diagram of a data processor 320. The encoded data for each channel data stream (i.e., the encoded data stream, X) is provided to a respective channel data processor 332. If the channel data stream is to be transmitted over multiple sub-channels and/or multiple antennas (without duplication on at least some of the transmissions), channel data processor 332 demultiplexes the channel data stream into a number of (up to $L \bullet N_T$) data sub-streams. Each data sub-stream corresponds to a transmission on a particular sub-channel at a particular antenna. In typical implementations, the number of data sub-streams is less than $L \bullet N_T$ since some of the sub-channels are used for signaling, voice, and other types of data. The data sub-streams are then processed to generate corresponding sub-streams for each of the assigned sub-channels that are then provided to combiners 334. Combiners 334 combine the modulation symbols designated for each antenna into modulation symbol vectors that are then provided as a modulation symbol vector stream. The N_T modulation symbol vector streams for the N_T antennas are then provided to the subsequent processing blocks (i.e., modulators 114).

In a design that provides the most flexibility, best performance, and highest efficiency, the modulation symbol to be transmitted at each time slot, on each sub-channel, can be individually and independently selected. This feature allows for the best use of the available resource over all three dimensions - time, frequency, and space. The number of data bits transmitted by each modulation symbol may thus differ.

FIG. 4A is a block diagram of a channel data processor 400 that can be used for processing one channel data steam. Channel data processor 400 can

be used to implement one channel data processor 332 in FIG. 3. The transmission of a channel data stream may occur on multiple sub-channels (e.g., as for data 1 in FIG. 2) and may also occur from multiple antennas. The transmission on each sub-channel and from each antenna can represent non-duplicated data.

Within channel data processor 400, a demultiplexer 420 receives and demultiplexes the encoded data stream, X_i, into a number of sub-channel data streams, $X_{i,1}$ through $X_{i,M}$, one sub-channel data stream for each sub-channel being used to transmit data. The data demultiplexing can be uniform or non-uniform. For example, if some information about the transmission paths is known (i.e., full CSI or partial CSI is known), demultiplexer 420 may direct more data bits to the sub-channels capable of transmitting more bps/Hz. However, if no CSI is known, demultiplexer 420 may uniformly directs approximately equal numbers of bits to each of the allocated sub-channels.

Each sub-channel data stream is then provided to a respective spatial division processor 430. Each spatial division processor 430 may further demultiplex the received sub-channel data stream into a number of (up to N_T) data sub-streams, one data sub-stream for each antenna used to transmit the data. Thus, after demultiplexer 420 and spatial division processor 430, the encoded data stream X_i may be demultiplexed into up to $L{\bullet}N_T$ data sub-streams to be transmitted on up to L sub-channels from up to N_T antennas.

At any particular time slot, up to N_T modulation symbols may be generated by each spatial division processor 430 and provided to N_T combiners 400a through 440t. For example, spatial division processor 430a assigned to sub-channel 1 may provide up to N_T modulation symbols for sub-channel 1 of antennas 1 through N_T. Similarly, spatial division processor 430k assigned to sub-channel k may provide up to N_T symbols for sub-channel k of antennas 1 through N_T. Each combiner 440 receives the modulation symbols for the L sub-channels, combines the symbols for each time slot into a modulation symbol vector, and provides the modulation

symbol vectors as a modulation symbol vector stream, V, to the next processing stage (e.g., modulator 114).

Channel data processor 400 may also be designed to provide the necessary processing to implement the full-CSI or partial-CSI processing modes described above. The CSI processing may be performed based on the available CSI information and on selected channel data streams, sub-channels, antennas, etc. The CSI processing may also be enabled and disabled selectively and dynamically. For example, the CSI processing may be enabled for a particular transmission and disabled for some other transmissions. The CSI processing may be enabled under certain conditions, for example, when the transmission link has adequate C/I.

Channel data processor 400 in FIG. 4A provides a high level of flexibility. However, such flexibility is typically not needed for all channel data streams. For example, the data for a voice call is typically transmitted over one sub-channel for the duration of the call, or until such time as the sub-channel is reassigned. The design of the channel data processor can be greatly simplified for these channel data streams.

FIG. 4B is a block diagram of the processing that can be employed for one channel data steam such as overhead data, signaling, voice, or traffic data. A spatial division processor 450 can be used to implement one channel data processor 332 in FIG. 3 and can be used to support a channel data stream such as, for example, a voice call. A voice call is typically assigned to one sub-channel for multiple time slots (e.g., voice 1 in FIG. 2) and may be transmitted from multiple antennas. The encoded data stream, X_i, is provided to spatial division processor 450 that groups the data into blocks, with each block having a particular number of bits that are used to generate a modulation symbol. The modulation symbols from spatial division processor 450 are then provided to one or more combiners 440 associated with the one or more antennas used to transmit the channel data stream.

A specific implementation of a transmitter unit capable of generating the transmit signal shown in FIG. 2 is now described for a better

understanding of the invention. At time slot 2 in FIG. 2, control data is transmitted on sub-channel 1, broadcast data is transmitted on sub-channel 2, voice calls 1 and 2 are assigned to sub-channels 3 and 4, respectively, and traffic data is transmitted on sub-channels 5 through 16. In this example, the transmitter unit is assumed to include four transmit antennas (i.e., $N_T = 4$) and four transmit signals (i.e., four RF modulated signals) are generated for the four antennas.

FIG. 5A is a block diagram of a portion of the processing units that can be used to generate the transmit signal for time slot 2 in FIG. 2. The input data stream is provided to a demultiplexer (DEMUX) 510 that demultiplexes the stream into five channel data streams, S_1 through S_5, corresponding to control, broadcast, voice 1, voice 2, and data 1 in FIG. 2. Each channel data stream is provided to a respective encoder 512 that encodes the data using an encoding scheme selected for that stream.

In this example, channel data streams S_1 through S_3 are transmitted using transmit diversity. Thus, each of the encoded data streams X_1 through X_3 is provided to a respective channel data processor 532 that generates the modulation symbols for that stream. The modulation symbols from each of the channel data processors 532a through 532c are then provided to all four combiners 540a through 540d. Each combiner 540 receives the modulation symbols for all 16 sub-channels designated for the antenna associated with the combiner, combines the symbols on each sub-channel at each time slot to generate a modulation symbol vector, and provides the modulation symbol vectors as a modulation symbol vector stream, V, to an associated modulator 114. As indicated in FIG. 5A, channel data stream S_1 is transmitted on sub-channel 1 from all four antennas, channel data stream S_2 is transmitted on sub-channel 2 from all four antennas, and channel data stream S_3 is transmitted on sub-channel 3 from all four antennas.

FIG. 5B is a block diagram of a portion of the processing units used to process the encoded data for channel data stream S_4. In this example, channel data stream S_4 is transmitted using spatial diversity (and not transmit diversity as used for channel data streams S_1 through S_3). With

spatial diversity, data is demultiplexed and transmitted (concurrently in each of the assigned sub-channels or over different time slots) over multiple antennas. The encoded data stream X_4 is provided to a channel data processor 532d that generates the modulation symbols for that stream. The modulation symbols in this case are linear combinations of modulation symbols selected from symbol alphabets that correspond to each of the eigenmodes of the channel. In this example, there are four distinct eigenmodes, each of which is capable of conveying a different amount of information. As an example, suppose eigenmode 1 has a C/I that allows 64-QAM (6 bits) to be transmitted reliably, eigenmode 2 permits 16-QAM (4 bits), eigenmode 3 permits QPSK (2 bits) and eigenmode 4 permits BPSK (1 bit) to be used. Thus, the combination of all four eigenmodes allows a total of 13 information bits to be transmitted simultaneously as an effective modulation symbol on all four antennas in the same sub-channel. The effective modulation symbol for the assigned sub-channel on each antenna is a linear combination of the individual symbols associated with each eigenmode, as described by the matrix multiply given in equation (1) above.

FIG. 5C is a block diagram of a portion of the processing units used to process channel data stream S_5. The encoded data stream X_5 is provided to a demultiplexer (DEMUX) 530 that demultiplexes the stream X_5 into twelve sub-channel data streams, $X_{5,11}$ through $X_{5,16}$, one sub-channel data stream for each of the allocated sub-channels 5 through 16. Each sub-channel data stream is then provided to a respective sub-channel data processor 536 that generates the modulation symbols for the associated sub-channel data stream. The sub-channel symbol stream from sub-channel data processors 536a through 536l are then provided to demultiplexers 538a through 538l, respectively. Each demultiplexer 538 demultiplexes the received sub-channel symbol stream into four symbol sub-streams, with each symbol sub-stream corresponding to a particular sub-channel at a particular antenna. The four symbol sub-streams from each demultiplexer 538 are then provided to the four combiners 540a through 540d.

In FIG. 5C, a sub-channel data stream is processed to generate a sub-channel symbol stream that is then demultiplexed into four symbol sub-streams, one symbol sub-stream for a particular sub-channel of each antenna. This implementation is a different from that described for FIG. 4A. In FIG. 4A, the sub-channel data stream designated for a particular sub-channel is demultiplexed into a number of data sub-streams, one data sub-stream for each antenna, and then processed to generate the corresponding symbol sub-streams. The demultiplexing in FIG. 5C is performed after the symbol modulation whereas the demultiplexing in FIG. 4A is performed before the symbol modulation. Other implementations may also be used and are within the scope of the present invention.

Each combination of sub-channel data processor 536 and demultiplexer 538 in FIG. 5C performs in similar manner as the combination of sub-channel data processor 532d and demultiplexer 534d in FIG. 5B. The rate of each symbol sub-stream from each demultiplexer 538 is, on the average, a quarter of the rate of the symbol stream from the associated channel data processor 536.

FIG. 6 is a block diagram of a receiver unit 600, having multiple receive antennas, which can be used to receive one or more channel data streams. One or more transmitted signals from one or more transmit antennas can be received by each of antennas 610a through 610r and routed to a respective front end processor 612. For example, receive antenna 610a may receive a number of transmitted signals from a number of transmit antennas, and receive antenna 610r may similarly receive multiple transmitted signals. Each front end processor 612 conditions (e.g., filters and amplifies) the received signal, downconverts the conditioned signal to an intermediate frequency or baseband, and samples and quantizes the downconverted signal. Each front end processor 612 typically further demodulates the samples associated with the specific antenna with the received pilot to generate "coherent" samples that are then provided to a respective FFT processor 614, one for each receive antenna.

Each FFT processor 614 generates transformed representations of the received samples and provides a respective stream of modulation symbol vectors. The modulation symbol vector streams from FFT processors 614a through 614r are then provided to demultiplexer and combiners 620, which channelizes the stream of modulation symbol vectors from each FFT processor 614 into a number of (up to L) sub-channel symbol streams. The sub-channel symbol streams from all FFT processors 614 are then processed, based on the (e.g., diversity or MIMO) communications mode used, prior to demodulation and decoding.

For a channel data stream transmitted using the diversity communications mode, the sub-channel symbol streams from all antennas used for the transmission of the channel data stream are presented to a combiner that combines the redundant information across time, space, and frequency. The stream of combined modulation symbols are then provided to a (diversity) channel processor 630 and demodulated accordingly.

For a channel data stream transmitted using the MIMO communications mode, all sub-channel symbol streams used for the transmission of the channel data stream are presented to a MIMO processor that orthogonalizes the received modulation symbols in each sub-channel into the distinct eigenmodes. The MIMO processor performs the processing described by equation (2) above and generates a number of independent symbol sub-streams corresponding to the number of eigenmodes used at the transmitter unit. For example, MIMO processor can perform multiplication of the received modulation symbols with the left eigenvectors to generate post-conditioned modulation symbols, which correspond to the modulation symbols prior to the full-CSI processor at the transmitter unit. The (post-conditioned) symbol sub-streams are then provided to a (MIMO) channel processor 630 and demodulated accordingly. Thus, each channel processor 630 receives a stream of modulation symbols (for the diversity communications mode) or a number of symbol sub-streams (for the MIMO communications mode). Each stream or sub-stream of modulation symbols is then provided to a respective demodulator (DEMOD) that implements a

demodulation scheme (e.g., M-PSK, M-QAM, or others) that is complementary to the modulation scheme used at the transmitter unit for the sub-channel being processed. For the MIMO communications mode, the demodulated data from all assigned demodulators may then be decoded independently or multiplexed into one channel data stream and then decoded, depending upon the coding and modulation method employed at the transmitter unit. For both the diversity and MIMO communications modes, the channel data stream from channel processor 630 may then provided to a respective decoder 640 that implements a decoding scheme complementary to that used at the transmitter unit for the channel data stream. The decoded data from each decoder 540 represents an estimate of the transmitted data for that channel data stream.

FIG. 6 represents one embodiment of a receiver unit. Other designs can contemplated and are within the scope of the present invention. For example, a receiver unit may be designed with only one receive antenna, or may be designed capable of simultaneously processing multiple (e.g., voice, data) channel data streams.

As noted above, multi-carrier modulation is used in the communications system of the invention. In particular, OFDM modulation can be employed to provide a number of benefits including improved performance in a multipath environment, reduced implementation complexity (in a relative sense, for the MIMO mode of operation), and flexibility. However, other variants of multi-carrier modulation can also be used and are within the scope of the present invention.

OFDM modulation can improve system performance due to multipath delay spread or differential path delay introduced by the propagation environment between the transmitting antenna and the receiver antenna. The communications link (i.e., the RF channel) has a delay spread that may potentially be greater than the reciprocal of the system operating bandwidth, W. Because of this, a communications system employing a modulation scheme that has a transmit symbol duration of less than the delay spread will experience inter-symbol interference (ISI). The ISI

distorts the received symbol and increases the likelihood of incorrect detection.

With OFDM modulation, the transmission channel (or operating bandwidth) is essentially divided into a (large) number of parallel sub-channels (or sub-bands) that are used to communicate the data. Because each of the sub-channels has a bandwidth that is typically much less than the coherence bandwidth of the communications link, ISI due to delay spread in the link is significantly reduced or eliminated using OFDM modulation. In contrast, most conventional modulation schemes (e.g., QPSK) are sensitive to ISI unless the transmission symbol rate is small compared to the delay spread of the communications link.

As noted above, cyclic prefixes can be used to combat the deleterious effects of multipath. A cyclic prefix is a portion of an OFDM symbol (usually the front portion, after the IFFT) that is wrapped around to the back of the symbol. The cyclic prefix is used to retain orthogonality of the OFDM symbol, which is typically destroyed by multipath.

As an example, consider a communications system in which the channel delay spread is less than 10 μsec. Each OFDM symbol has appended onto it a cyclic prefix that insures that the overall symbol retains its orthogonal properties in the presence of multipath delay spread. Since the cyclic prefix conveys no additional information, it is essentially overhead. To maintain good efficiency, the duration of the cyclic prefix is selected to be a small fraction of the overall transmission symbol duration. For the above example, using a 5% overhead to account for the cyclic prefix, a transmission symbol duration of 200 μsec is adequate for a 10 μsec maximum channel delay spread. The 200 μsec transmission symbol duration corresponds to a bandwidth of 5 kHz for each of the sub-bands. If the overall system bandwidth is 1.2288 MHz, 250 sub-channels of approximately 5 kHz can be provided. In practice, it is convenient for the number of sub-channels to be a power of two. Thus, if the transmission symbol duration is increased to

205 μsec and the system bandwidth is divided into M = 256 sub-bands, each sub-channel will have a bandwidth of 4.88 kHz.

In certain embodiments of the invention, OFDM modulation can reduce the complexity of the system. When the communications system incorporates MIMO technology, the complexity associated with the receiver unit can be significant, particularly when multipath is present. The use of OFDM modulation allows each of the sub-channels to be treated in an independent manner by the MIMO processing employed. Thus, OFDM modulation can significantly simplify the signal processing at the receiver unit when MIMO technology is used.

OFDM modulation can also afford added flexibility in sharing the system bandwidth, W, among multiple users. Specifically, the available transmission space for OFDM symbols can be shared among a group of users. For example, low rate voice users can be allocated a sub-channel or a fraction of a sub-channel in OFDM symbol, while the remaining sub-channels can be allocated to data users based on aggregate demand. In addition, overhead, broadcast, and control data can be conveyed in some of the available sub-channels or (possibly) in a portion of a sub-channel.

As described above, each sub-channel at each time slot is associated with a modulation symbol that is selected from some alphabet such as M-PSK or M-QAM. In certain embodiments, the modulation symbol in each of the L sub-channels can be selected such that the most efficient use is made of that sub-channel. For example, sub-channel 1 can be generated using QPSK, sub-channel 2 can be generate using BPSK, sub-channel 3 can be generated using 16-QAM, and so on. Thus, for each time slot, up to L modulation symbols for the L sub-channels are generated and combined to generate the modulation symbol vector for that time slot.

One or more sub-channels can be allocated to one or more users. For example, each voice user may be allocated a single sub-channel. The remaining sub-channels can be dynamically allocated to data users. In this case, the remaining sub-channels can be allocated to a single data user or divided among multiple data users. In addition, some sub-channels can be

reserved for transmitting overhead, broadcast, and control data. In certain embodiments of the invention, it may be desirable to change the sub-channel assignment from (possibly) modulation symbol to symbol in a pseudo-random manner to increase diversity and provide some interference averaging.

In a CDMA system, the transmit power on each reverse link transmission is controlled such that the required frame error rate (FER) is achieved at the base station at the minimal transmit power, thereby minimizing interference to other users in the system. On the forward link of the CDMA system, the transmit power is also adjusted to increase system capacity.

In the communications system of the invention, the transmit power on the forward and reverse links can be controlled to minimize interference and maximize system capacity. Power control can be achieved in various manners. For example, power control can be performed on each channel data stream, on each sub-channel, on each antenna, or on some other unit of measurement. When operating in the diversity communications mode, if the path loss from a particular antenna is great, transmission from this antenna can be reduced or muted since little may be gained at the receiver unit. Similarly, if transmission occurs over multiple sub-channels, less power may be transmitted on the sub-channel(s) experiencing the most path loss.

In an implementation, power control can be achieved with a feedback mechanism similar to that used in the CDMA system. Power control information can be sent periodically or autonomously from the receiver unit to the transmitter unit to direct the transmitter unit to increase or decrease its transmit power. The power control bits may be generated based on, for example, the BER or FER at the receiver unit.

FIG. 7 shows plots that illustrate the spectral efficiency associated with some of the communications modes of the communications system of the invention. In FIG. 7, the number of bits per modulation symbol for a given bit error rate is given as a function of C/I for a number of system

configurations. The notation $N_T x N_R$ denotes the dimensionality of the configuration, with N_T = number of transmit antennas and N_R = number of receive antennas. Two diversity configurations, namely 1x2 and 1x4, and four MIMO configurations, namely 2x2, 2x4, 4x4, and 8x4, are simulated and the results are provided in FIG. 7.

As shown in the plots, the number of bits per symbol for a given BER ranges from less than 1 bps/Hz to almost 20 bps/Hz. At low values of C/I, the spectral efficiency of the diversity communications mode and MIMO communications mode is similar, and the improvement in efficiency is less noticeable. However, at higher values of C/I, the increase in spectral efficiency with the use of the MIMO communications mode becomes more dramatic. In certain MIMO configurations and for certain conditions, the instantaneous improvement can reach up to 20 times.

From these plots, it can be observed that spectral efficiency generally increases as the number of transmit and receive antennas increases. The improvement is also generally limited to the lower of N_T and N_R. For example, the diversity configurations, 1x2 and 1x4, both asymptotically reach approximately 6 bps/Hz.

In examining the various data rates achievable, the spectral efficiency values given in FIG. 7 can be applied to the results on a sub-channel basis to obtain the range of data rates possible for the sub-channel. As an example, for a subscriber unit operating at a C/I of 5 dB, the spectral efficiency achievable for this subscriber unit is between 1 bps/Hz and 2.25 bps/Hz, depending on the communications mode employed. Thus, in a 5 kHz sub-channel, this subscriber unit can sustain a peak data rate in the range of 5 kbps to 10.5 kbps. If the C/I is 10 dB, the same subscriber unit can sustain peak data rates in the range of 10.5 kbps to 25 kbps per sub-channel. With 256 sub-channels available, the peak sustained data rate for a subscriber unit operating at 10 dB C/I is then 6.4 Mbps. Thus, given the data rate requirements of the subscriber unit and the operating C/I for the subscriber unit, the system can allocate the necessary number of sub-channels to meet the requirements. In the case of data services, the number of sub-channels

allocated per time slot may vary depending on, for example, other traffic loading.

The reverse link of the communications system can be designed similar in structure to the forward link. However, instead of broadcast and common control channels, there may be random access channels defined in specific sub-channels or in specific modulation symbol positions of the frame, or both. These may be used by some or all subscriber units to send short requests (e.g., registration, request for resources, and so on) to the central station. In the common access channels, the subscriber units may employ common modulation and coding. The remaining channels may be allocated to separate users as in the forward link. Allocation and de-allocation of resources (on both the forward and reverse links) can be controlled by the system and can be communicated on the control channel in the forward link.

One design consideration for on the reverse link is the maximum differential propagation delay between the closest subscriber unit and the furthest subscriber unit. In systems where this delay is small relative to the cyclic prefix duration, it may not be necessary to perform correction at the transmitter unit. However, in systems in which the delay is significant, the cyclic prefix can be extended to account for the incremental delay. In some instances, it may be possible to make a reasonable estimate of the round trip delay and correct the time of transmit so that the symbol arrives at the central station at the correct instant. Usually there is some residual error, so the cyclic prefix may also further be extended to accommodate this residual error.

In the communications system, some subscriber units in the coverage area may be able to receive signals from more than one central station. If the information transmitted by multiple central stations is redundant on two or more sub-channels and/or from two or more antennas, the received signals can be combined and demodulated by the subscriber unit using a diversity-combining scheme. If the cyclic prefix employed is sufficient to handle the differential propagation delay between the earliest and latest

arrival, the signals can be (optimally) combined in the receiver and demodulated correctly. This diversity reception is well known in broadcast applications of OFDM. When the sub-channels are allocated to specific subscriber units, it is possible for the same information on a specific sub-channel to be transmitted from a number of central stations to a specific subscriber unit. This concept is similar to the soft handoff used in CDMA systems.

As shown above, the transmitter unit and receiver unit are each implemented with various processing units that include various types of data processor, encoders, IFFTs, FFTs, demultiplexers, combiners, and so on. These processing units can be implemented in various manners such as an application specific integrated circuit (ASIC), a digital signal processor, a microcontroller, a microprocessor, or other electronic circuits designed to perform the functions described herein. Also, the processing units can be implemented with a general-purpose processor or a specially designed processor operated to execute instruction codes that achieve the functions described herein. Thus, the processing units described herein can be implemented using hardware, software, or a combination thereof.

The foregoing description of the preferred embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without the use of the inventive faculty. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

WHAT IS CLAIMED IS:

CLAIMS

1. A method for measuring and reporting transmission characteristics of a propagation channel in a multiple input/multiple output communication system, comprising the steps of:

generating a plurality of pilot signals;

transmitting the plurality of pilot signals over a propagation channel between a transmitter unit and a plurality of receiver units, wherein the transmitter unit comprises at least one transmit antenna, each of the plurality of receiver units comprises at least one receive antenna, and the propagation channel comprises a plurality of sub-channels between the transmitter unit and the plurality of receiver units;

receiving at least one of the plurality of pilot signals at each of the plurality of receiver units;

determining a set of transmission characteristics for at least one of the plurality of sub-channels, wherein the step of determining the set of transmission characteristics uses at least one of the plurality of pilot signals received at each of the plurality of receiver units;

reporting an information signal from each of the plurality of receiver units to the transmitter unit, wherein the information signal carries the set of transmission characteristics for at least one of the plurality of sub-channels; and

optimizing a set of transmission parameters at the transmitter unit, based on the information signal.

2. The method of Claim 1, wherein the step of transmitting the plurality of pilot signals comprise the steps of:

generating a plurality of disjoint orthogonal frequency-division multiplexed (OFDM) sub-channel sets, wherein the plurality of disjoint OFDM sub-channel sets may comprise disjoint, substantially orthogonal frequency-division multiplexed sub-channel sets; and

transmitting at least one of the plurality of pilot signals on at least one of the plurality of disjoint OFDM sub-channel sets.

3. The method of Claim 2, wherein the step of generating the plurality of disjoint OFDM sub-channel sets comprises the step of reusing at least one of the plurality of disjoint OFDM sub-channel sets if the at least one transmit antenna is spatially distant from any other transmit antenna.

4. The method of Claim 2, wherein the step of determining the set of transmission characteristics for at least one of the plurality of sub-channels comprises the step of analyzing a group of the disjoint OFDM sub-channel sets.

5. The method of Claim 4, wherein the set of transmission characteristics comprises an average interference level.

6. The method of Claim 4, wherein the set of transmission characteristics comprises a noise level.

7. The method of Claim 1, wherein the plurality of pilot signals comprises a plurality of orthogonal sequences.

8. The method of Claim 1, wherein the plurality of pilot signals comprises a plurality of OFDM symbols.

9. The method of Claim 8, wherein the plurality of OFDM symbols are orthogonally coded.

10. The method of Claim 9, wherein the plurality of OFDM symbols are orthogonally coded with Walsh code sequences.

11. The method of Claim 1, wherein the plurality of pilot signals comprises a plurality of shifted Maximal-Length Shift Register sequences (m-sequences), wherein each of the plurality of shifted m-sequences is separated by a predetermined period.

12. The method of Claim 11, wherein the plurality of pilot signals comprises a plurality of shifted, appended m-sequences, wherein each of the plurality of shifted, appended m-sequences includes a repeated portion of the m-sequence.

13. The method of Claim 4, wherein the group of the disjoint OFDM sub-channel sets comprises:

at least one of the plurality of disjoint OFDM sub-channel sets associated with a principal link; and

at least one of the plurality of disjoint OFDM sub-channel sets associated with a set of interfering links.

14. The method of Claim 13, wherein the information signal carries the set of transmission characteristics associated with the principal link and the set of interfering links.

15. The method of Claim 2, wherein the step of reporting transmission parameters comprises the steps of:

generating a polynomial function representative of a set of transmission characteristics of the principle link; and

transmitting a set of coefficients associated with the polynomial function.

16. The method of Claim 2, wherein the step of reporting the information signal comprises the step of compressing the set of transmission characteristics for at least one of the plurality of sub-channels, wherein the set of transmission characteristics is obtained from an inverse fast Fourier transformation performed upon a channel frequency response.

17. The method of Claim 1, further comprising the steps of:

generating a plurality of scheduling messages at the transmitter unit; and

transmitting at least one of the plurality of scheduling messages to at least one of the plurality of receiver units, wherein upon receipt of the at least one of the plurality of scheduling messages, the at least one of the plurality of receiver units schedules the step of reporting the information signal.

18. An apparatus for measuring and reporting transmission characteristics of a propagation channel in a multiple input/multiple output communication system, comprising:

means for generating a plurality of pilot signals;

means for transmitting the plurality of pilot signals over a propagation channel between a transmitter unit and a plurality of receiver units, wherein the transmitter unit comprises at least one transmit antenna, each of the plurality of receiver units comprises at least one receive antenna, and the propagation channel comprises a plurality of sub-channels between the transmitter unit and the plurality of receiver units;

means for receiving at least one of the plurality of pilot signals at the receiver unit;

means for determining a set of transmission characteristics for at least one of the plurality of sub-channels, wherein the step of determining the set of transmission characteristics uses at least one of the plurality of pilot signals received at each of the plurality of receiver units;

means for reporting an information signal from each of the plurality of receiver units to the transmitter unit, wherein the information signal carries the set of transmission characteristics for at least one of the plurality of sub-channels; and

means for optimizing a set of transmission parameters at the transmitter unit, based on the information signal.

19. A method for measuring and reporting channel state information (CSI) in a multiple input/multiple output (MIMO) system, comprising the steps of:

assigning a plurality of disjoint sub-channel sets to a plurality of transmit antennas;

transmitting a plurality of Orthogonal Frequency Division Multiplexed (OFDM) pilot signals from a transmitter unit to a plurality of receiver units, wherein each of the plurality of OFDM pilot signals is transmitted on at least one of the plurality of disjoint sub-channel sets;

demodulating the plurality of OFDM pilot signals;

determining the CSI of the plurality of disjoint sub-channel sets, wherein the step of determining the CSI uses the demodulated plurality of OFDM pilot signals;

transmitting the CSI of the plurality of disjoint sub-channel sets to the transmitter unit; and

preconditioning a transmission symbol.

20. The method of Claim 19, wherein the step of transmitting the CSI of the plurality of disjoint sub-channel sets comprises the steps of:

compressing the CSI into a reduced matrix; and

transmitting a representation of the reduced matrix to the transmitter unit.

21. The method of Claim 20, wherein the reduced matrix is a multiplication result from multiplying a channel response matrix and a complex-conjugate of the channel response matrix, wherein the channel response matrix includes a plurality of the CSI gain values.

22. The method of Claim 21, wherein the representation of the reduced matrix is an eigenmode matrix.

22. The method of Claim 19, wherein the step of determining the CSI of the plurality of disjoint sub-channel sets further comprises the steps of:

determining whether a communications link has a number of multipath components that is less than a predefined threshold; and

performing a inverse Fast Fourier Transform (IFFT) operation on a set of channel frequency responses of the communications link if the number of multipath components is less than the predefined threshold, wherein the result of the IFFT operation is channel state information to be transmitted to the transmitter unit.

23. A system for measuring and reporting channel state information (CSI) in a multiple input/multiple output communication system, comprising:

a processor at a base station for assigning a plurality of disjoint sub-channel sets to a plurality of transmit antennas, for generating a plurality of pilot signals, for assigning each of the plurality of pilot signals to at least one of the plurality of disjoint sub-channel sets, and for preconditioning transmission data;

a modulator connected to the processor for receiving the plurality of pilot signals and modulating the plurality of pilot signals onto the plurality of assigned disjoint sub-channel sets, wherein the plurality of assigned disjoint sub-channel sets are transmitted by the plurality of transmit antennas;

a demodulator at each of a plurality of receiver units for receiving data carried on the plurality of disjoint sub-channel sets; and

a processor connected to the demodulator at each of the plurality of receiver units for analyzing demodulated data, wherein the processor determines CSI from demodulated data and generates a CSI message for transmission to the base station, wherein the CSI message is used by the processor at the base station to precondition transmission data.

24. The system of Claim 23, wherein the processor connected to the demodulator at each of the plurality of receiver units generates the CSI message for a subset of the plurality of disjoint sub-channel sets.

25. The system of Claim 23, wherein the processor at the base station generates a plurality of pilot signals that comprises a plurality of orthogonal sequences.

26. The system of Claim 23, wherein the processor at the base station generates a plurality of pilot signals that comprises a plurality of periodic OFDM symbols.

27. The system of Claim 23, wherein the processor at the base station generates a plurality of pilot signals that comprises a plurality of shifted Maximal-Length Shift Register sequences (m-sequences).

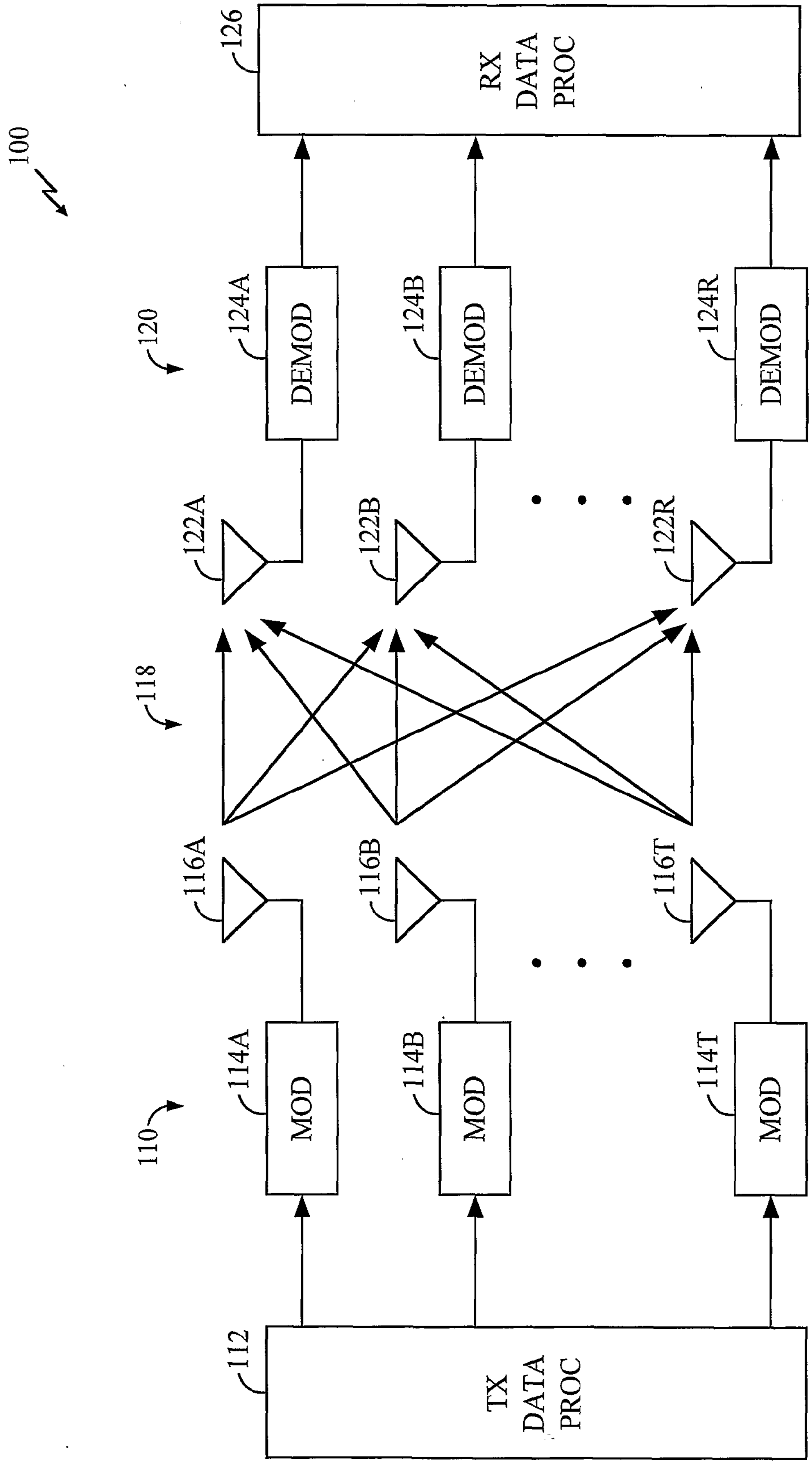

100
126
RX DATA PROC
120
124A
DEMOD
124B
DEMOD
124R
DEMOD
122A
122B
122R
118
116A
116B
116T
110
114A
MOD
114B
MOD
114T
MOD
112
TX DATA PROC

FIG. 1

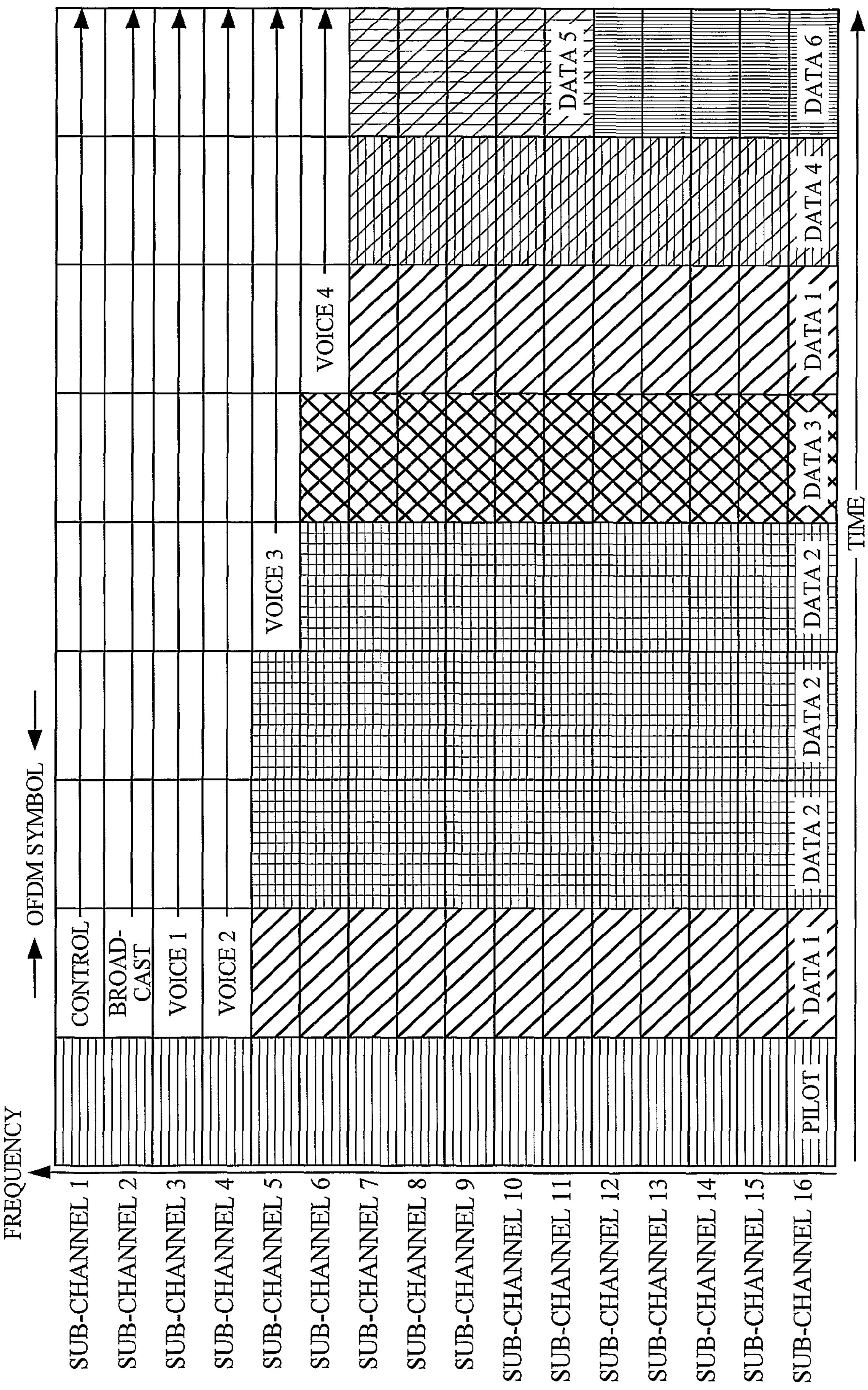

FREQUENCY
OFDM SYMBOL
SUB-CHANNEL 1
SUB-CHANNEL 2
SUB-CHANNEL 3
SUB-CHANNEL 4
SUB-CHANNEL 5
SUB-CHANNEL 6
SUB-CHANNEL 7
SUB-CHANNEL 8
SUB-CHANNEL 9
SUB-CHANNEL 10
SUB-CHANNEL 11
SUB-CHANNEL 12
SUB-CHANNEL 13
SUB-CHANNEL 14
SUB-CHANNEL 15
SUB-CHANNEL 16
CONTROL
BROAD-CAST
VOICE 1
VOICE 2
VOICE 3
VOICE 4
PILOT
DATA 1
DATA 2
DATA 2
DATA 2
DATA 3
DATA 1
DATA 4
DATA 5
DATA 6
TS1
TS2
TS3
TS4
TS5
TS6
TS7
TS8
TS9
TIME

FIG. 2

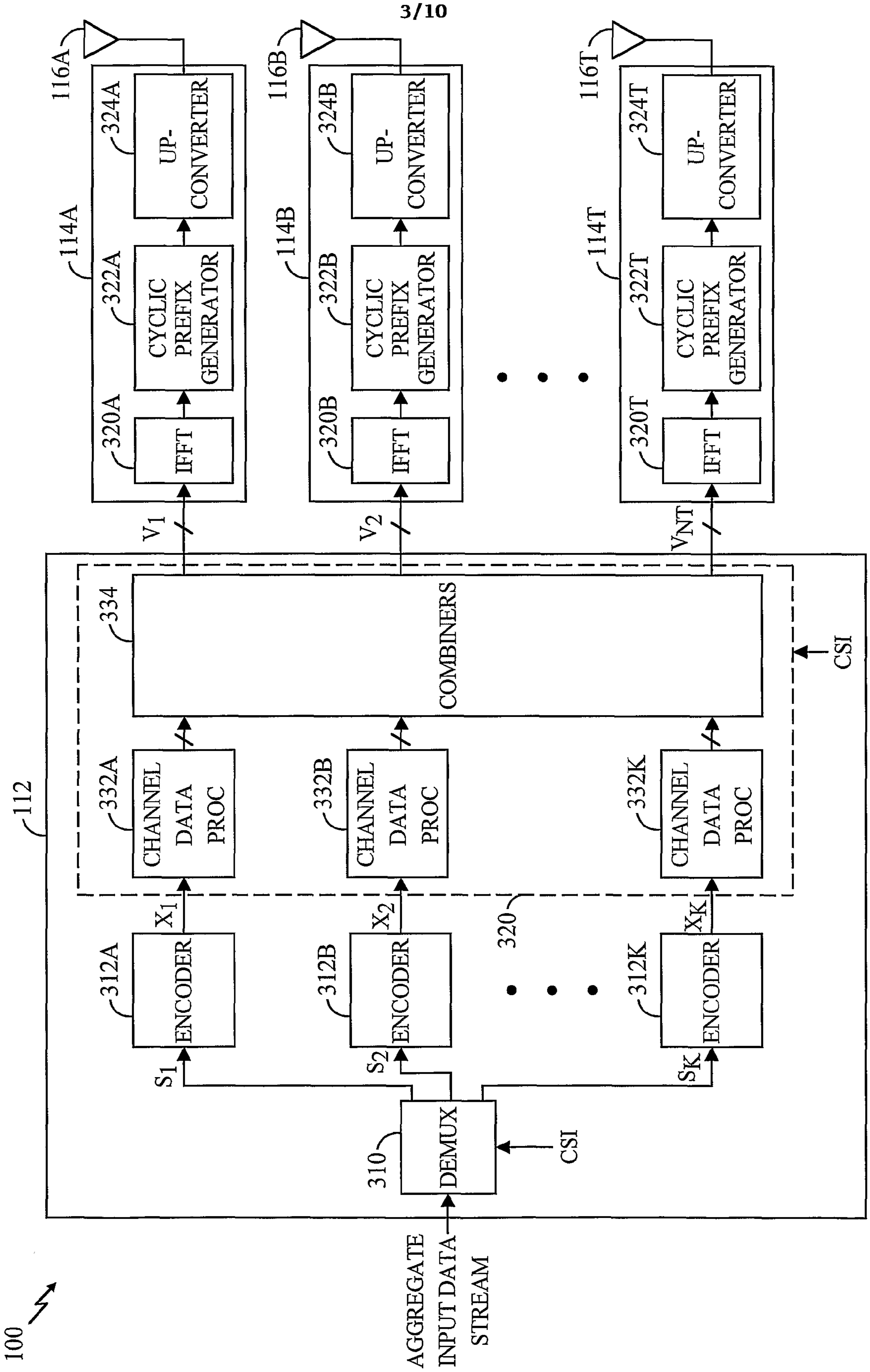

100
AGGREGATE INPUT DATA STREAM
112
310
DEMUX
CSI
S1
S2
SK
312A
ENCODER
312B
ENCODER
312K
ENCODER
X1
X2
XK
320
332A
CHANNEL DATA PROC
332B
CHANNEL DATA PROC
332K
CHANNEL DATA PROC
334
COMBINERS
CSI
V1
V2
VNT
114A
320A
IFFT
322A
CYCLIC PREFIX GENERATOR
324A
UP-CONVERTER
116A
114B
320B
IFFT
322B
CYCLIC PREFIX GENERATOR
324B
UP-CONVERTER
116B
114T
320T
IFFT
322T
CYCLIC PREFIX GENERATOR
324T
UP-CONVERTER
116T

FIG. 3

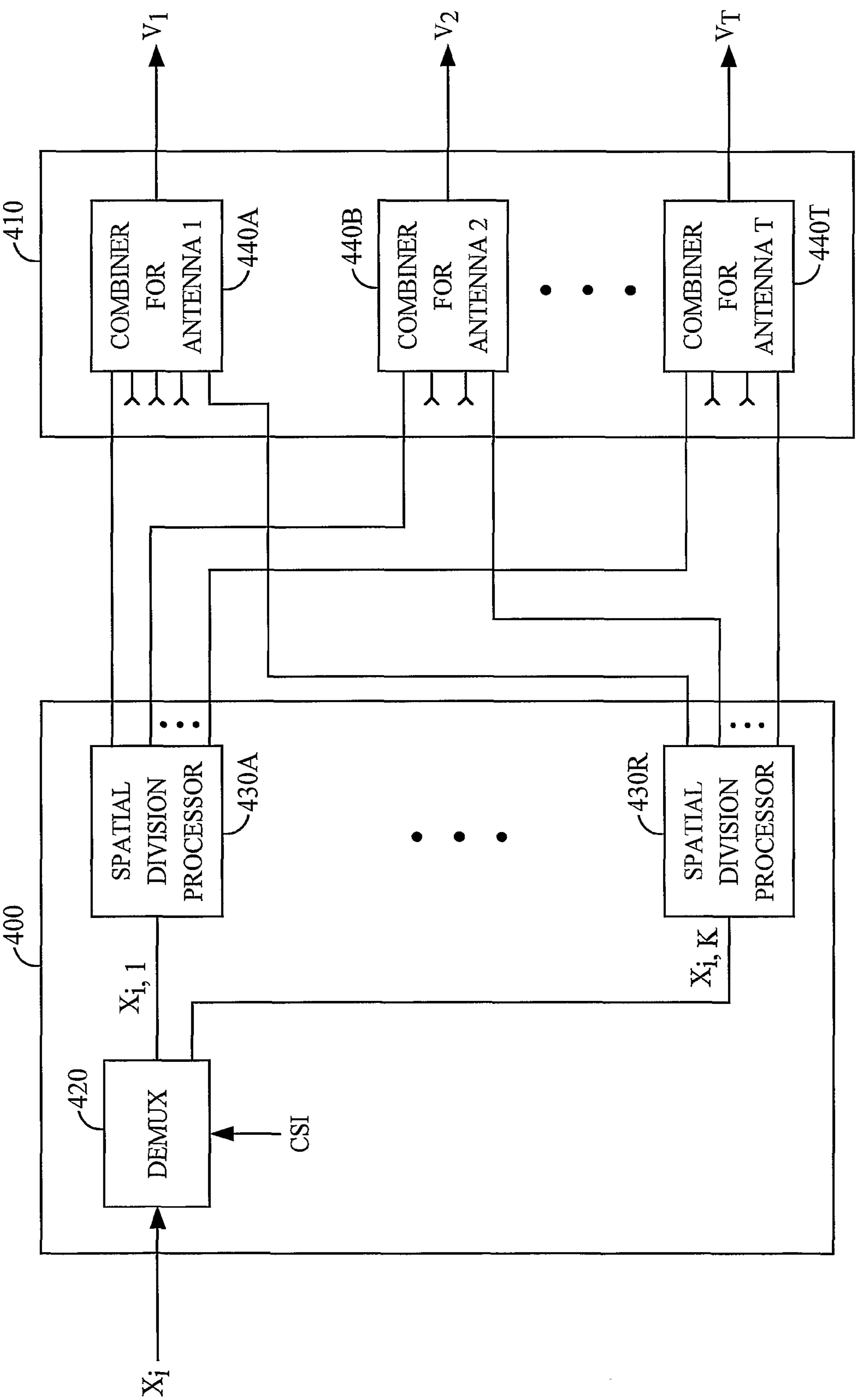
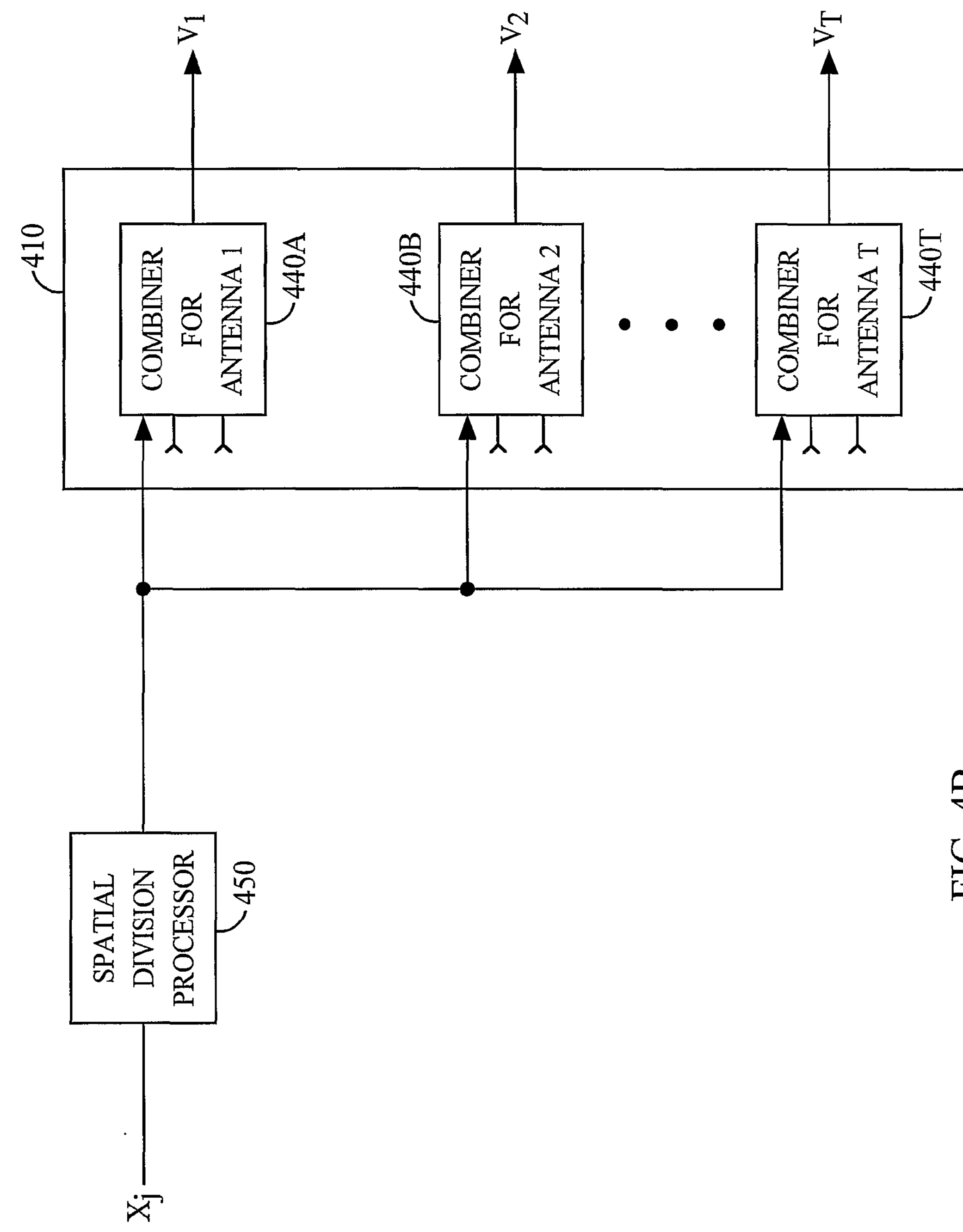
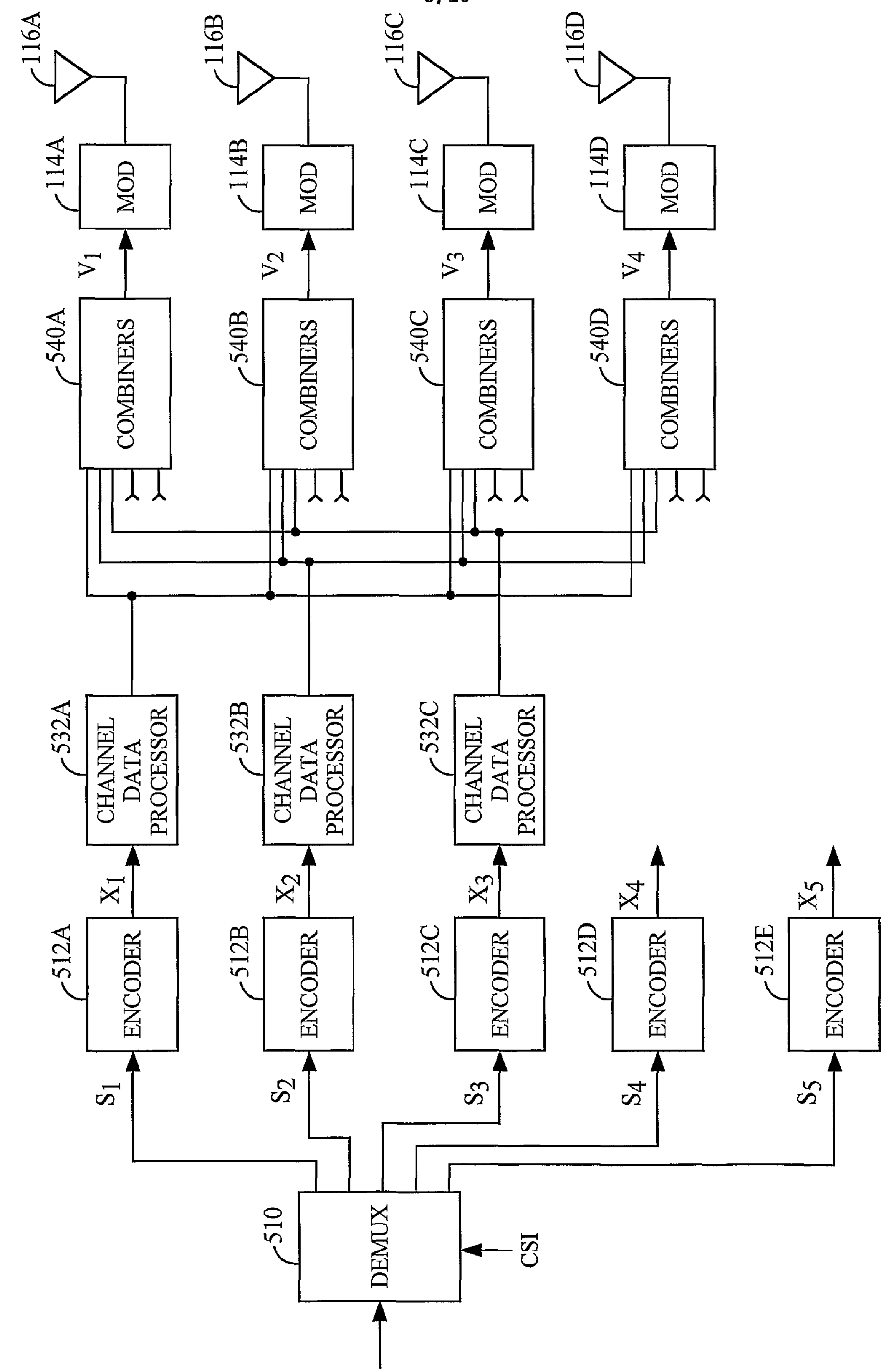

V_1
V_2
V_T
410
COMBINER FOR ANTENNA 1
440A
440B
COMBINER FOR ANTENNA 2
COMBINER FOR ANTENNA T
440T
400
SPATIAL DIVISION PROCESSOR
430A
430R
SPATIAL DIVISION PROCESSOR
$X_{i,1}$
$X_{i,K}$
420
DEMUX
CSI
X_i

FIG. 4A



V_1
V_2
V_T
410
COMBINER FOR ANTENNA 1
440A
440B
COMBINER FOR ANTENNA 2
COMBINER FOR ANTENNA T
440T
SPATIAL DIVISION PROCESSOR
450
X_j

FIG. 4B



510
DEMUX
CSI
S_1
S_2
S_3
S_4
S_5
512A
512B
512C
512D
512E
ENCODER
ENCODER
ENCODER
ENCODER
ENCODER
X_1
X_2
X_3
X_4
X_5
532A
532B
532C
CHANNEL DATA PROCESSOR
CHANNEL DATA PROCESSOR
CHANNEL DATA PROCESSOR
540A
540B
540C
540D
COMBINERS
COMBINERS
COMBINERS
COMBINERS
V_1
V_2
V_3
V_4
114A
114B
114C
114D
MOD
MOD
MOD
MOD
116A
116B
116C
116D

FIG. 5A

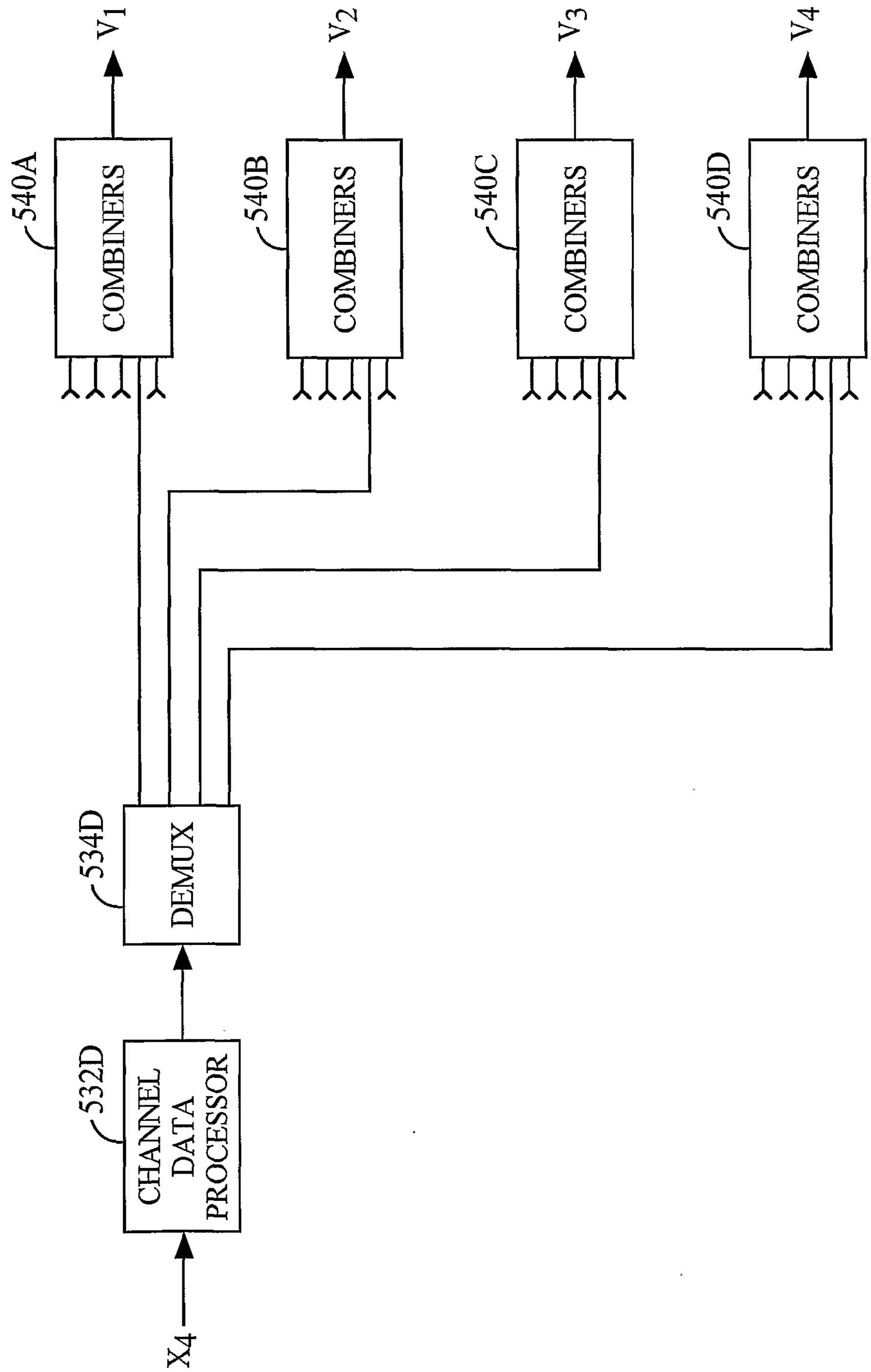
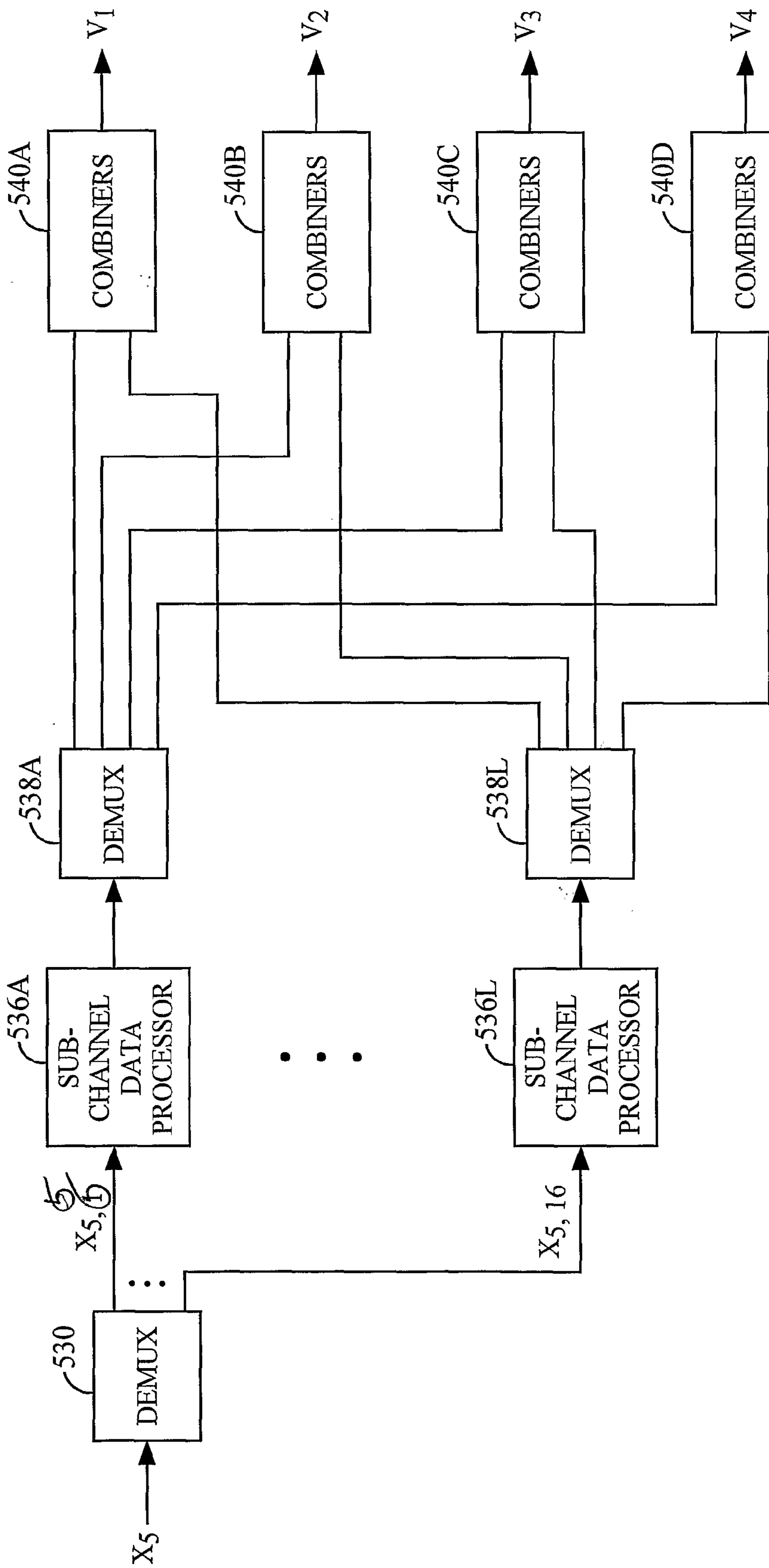

X4
532D
CHANNEL DATA PROCESSOR
534D
DEMUX
540A
COMBINERS
V1
540B
COMBINERS
V2
540C
COMBINERS
V3
540D
COMBINERS
V4

FIG. 5B



V1
V2
V3
V4
540A
540B
540C
540D
COMBINERS
COMBINERS
COMBINERS
COMBINERS
538A
DEMUX
538L
DEMUX
536A
SUB-CHANNEL DATA PROCESSOR
536L
SUB-CHANNEL DATA PROCESSOR
X5, 1
5
X5, 16
530
DEMUX
X5

FIG. 5C

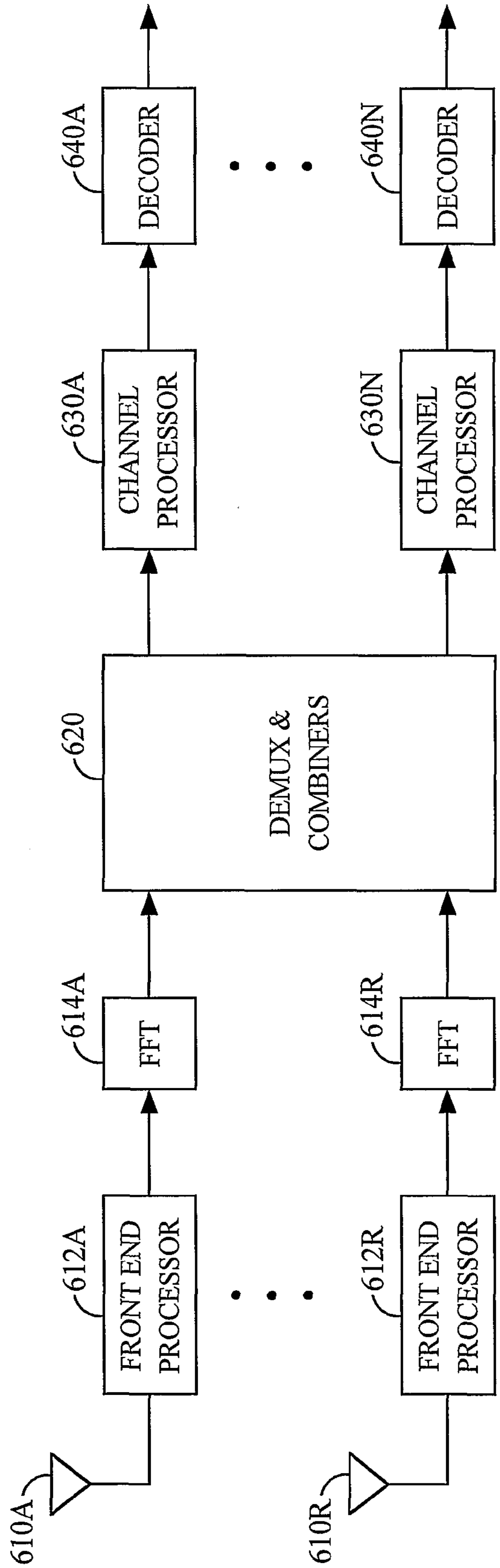

610A
612A
FRONT END PROCESSOR
614A
FFT
620
DEMUX & COMBINERS
630A
CHANNEL PROCESSOR
640A
DECODER
610R
612R
FRONT END PROCESSOR
614R
FFT
630N
CHANNEL PROCESSOR
640N
DECODER

FIG. 6

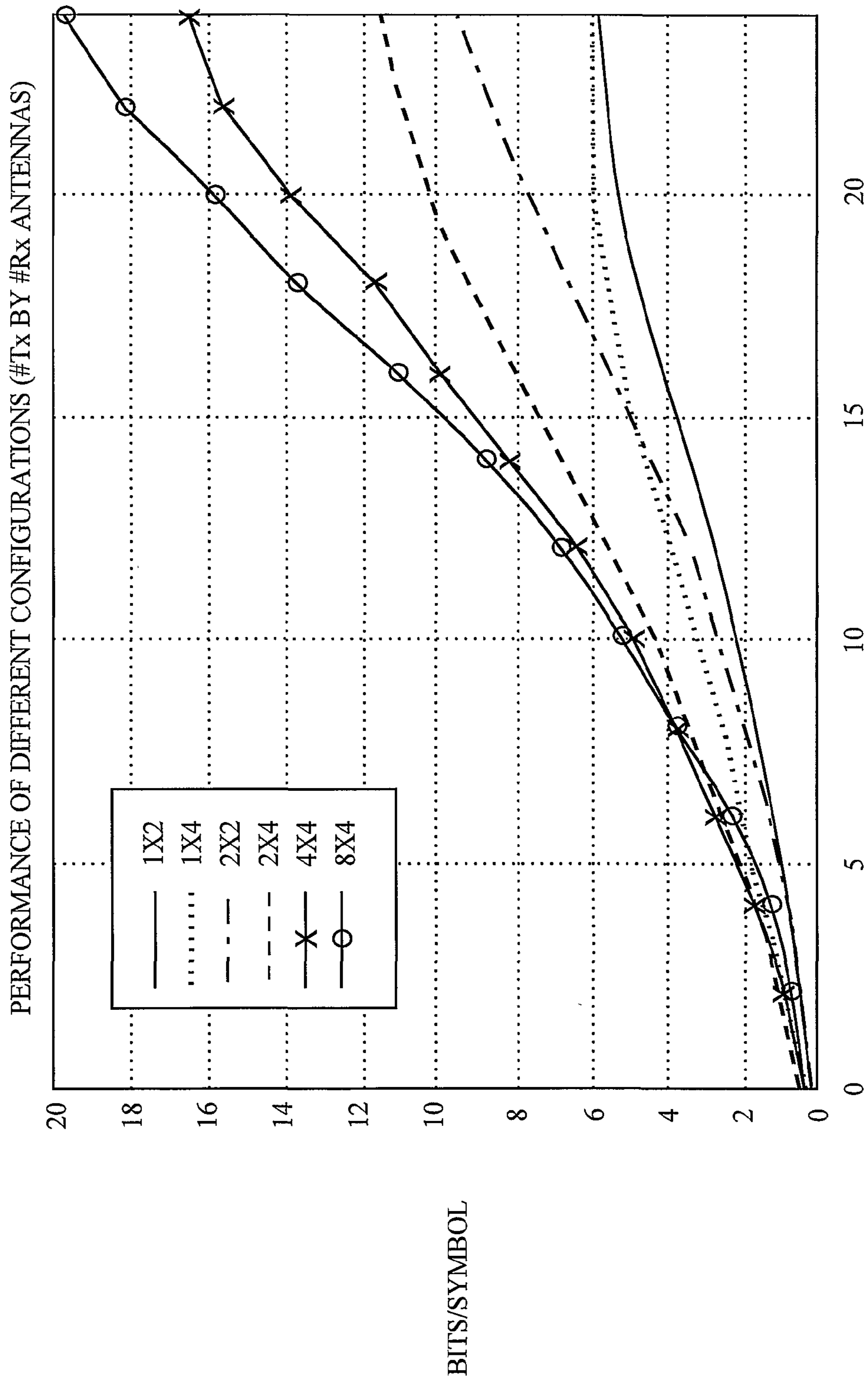

PERFORMANCE OF DIFFERENT CONFIGURATIONS (#Tx BY #Rx ANTENNAS)
1X2
1X4
2X2
2X4
4X4
8X4
BITS/SYMBOL
C/I (dB)
0
2
4
6
8
10
12
14
16
18
20
0
5
10
15
20

FIG. 7

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
25 October 2001 (25.10.2001)

PCT

(10) International Publication Number
WO 01/80510 A1

(51) International Patent Classification[7]: H04L 27/26

(21) International Application Number: PCT/US01/12555

(22) International Filing Date: 18 April 2001 (18.04.2001)

(25) Filing Language: English

(26) Publication Language: English

(30) Priority Data:
60/197,727 18 April 2000 (18.04.2000) US

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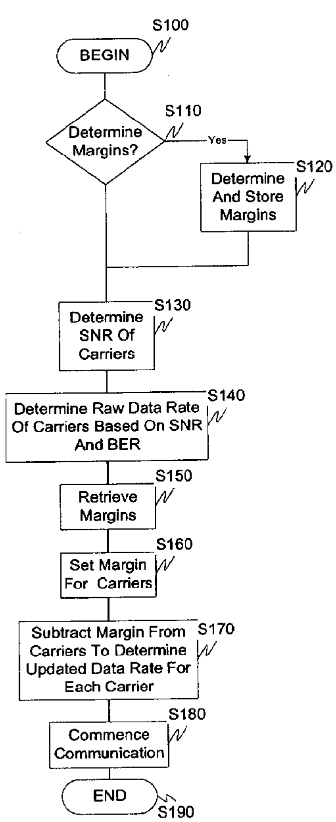
(81) Designated States *(national)*: AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, BZ, CA, CH, CN, CR, CU, CZ, DE, DK, DM, DZ, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NO, NZ, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TR, TT, TZ, UA, UG, UZ, VN, YU, ZA, ZW.

(84) Designated States *(regional)*: ARIPO patent (GH, GM, KE, LS, MW, MZ, SD, SL, SZ, TZ, UG, ZW), Eurasian patent (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European patent (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE, TR), OAPI patent (BF, BJ, CF, CG, CI, CM, GA, GN, GW, ML, MR, NE, SN, TD, TG).

Published:
— *with international search report*
— *before the expiration of the time limit for amending the claims and to be republished in the event of receipt of amendments*

[Continued on next page]

(54) Title: SYSTEMS AND METHODS FOR A MULTICARRIER MODULATION SYSTEM WITH A VARIABLE MARGIN



(57) Abstract: A multicarrier modem has a plurality of carriers over which data is transmitted. By assigning, for example, one or more different margins to the individual carriers, the data rate and impairment immunity can be increased.

WO 01/80510 A1

DATA ALLOCATION WITH VARIABLE SNR MARGINS

Background of the Invention

Field of the Invention

[0001] This invention relates to communications technologies. In particular, this invention relates to multicarrier modulation systems having multiple margins.

Description of Related Art

[0002] Multicarrier modulation, or Discrete Multitone Modulation (DMT), is a transmission method that is widely used for communication over difficult media. Multicarrier modulation divides the transmission frequency band into multiple subchannels, i.e., carriers or bins, with each carrier individually modulating a bit or a collection of bits. A transmitter modulates an input data stream containing information bits with one or more carriers, i.e., bins or subchannels, and transmits the modulated information. A receiver demodulates all the carriers in order to recover the transmitted information bits as an output data stream.

[0003] Multicarrier modulation has many advantages over single carrier modulation. These advantages include, for example, a higher immunity to impulse noise, a lower complexity equalization requirement in the presence of multipath, a higher immunity to narrow band interference, a higher data rate and bandwidth flexibility. Multicarrier modulation is being used in many applications to obtain these advantages, as well as for other reasons. These applications include Asymmetric Digital Subscriber Line (ADSL) systems, wireless LAN systems, power line communications systems, and other applications. ITU standards G.992.1 and G.992.2 and the ANSI T1.413 standard specify standard implementations for ADSL transceivers that use multicarrier modulation.

[0004] Discrete multitone modulation transceivers modulate a number of bits on each subchannel, the number of bits depending on the Signal to Noise Ratio (SNR) of that subchannel and the Bit Error Rate (BER) requirement of a link. For example, if the required BER is 1×10^{-7}, i.e., one bit in ten million is received in error on average, and the

SNR of a particular subchannel is 21.5 dB, then that subchannel can modulate 4 bits, since 21.5 dB is the required SNR to transmit 4 QAM bits with a 1×10^{-7} BER. Other subchannels can have a different SNR and therefore may have a different number of bits allocated to them at the same BER. Additional information regarding bit loading can be found in copending U.S. Application Serial No. 09/510,773, incorporated herein by reference in its entirety.

[0005] In many DMT systems, an additional parameter is used to determine the number of bits allocated to each subchannel. This parameter is called the SNR "margin," or simply the "margin." The margin specifies an extra SNR per subchannel, in addition to what is required to maintain the specified BER requirement. As an example, a DMT system with a 6 dB margin would require a 21.5+6=27.5 dB SNR on a subchannel in order to transmit 4 bits on that subchannel with a 1×10^{-7} BER. This is 6 dB more than required by the example in the previous paragraph because now a 6 dB margin is added to the system. Another way of looking at this is that in the example of the previous paragraph, where 4 bits were allocated to a subchannel with 21.5 dB SNR, the margin was 0 dB.

[0006] DMT transceivers use a margin to increase the system's immunity to various types of time varying impairments. Examples of these impairments in DSL systems are: changes in the levels of crosstalk from other transmission systems, impulse noise, temperature changes in the telephone line, or the like. When a DMT system is operating with a positive SNR margin, the noise can change instantaneously by the level of the margin and the system will still maintain the required BER. For example, if the system is operating at a 6 dB margin, e.g., 4 bits are allocated to carriers with 27.5 dB SNR for BER=1×10^{-7}, the crosstalk levels can increase by 6 dB and the system will still be operating at the required 1×10^{-7} BER. Obviously the penalty for this increase in robustness is a decrease in the data rate, since with a 0 dB margin, a subchannel with 27.5 dB SNR can modulate 6 bits at 1×10^{-7} BER.

[0007] Therefore, there is a tradeoff between the robustness of the channel, such as a phone line, and the achievable data rate. The margin can be used to quantify this tradeoff. A higher margin results in a higher level of immunity to changing channel conditions at the expense of the achievable data rate. Likewise, a lower margin results in a higher data rate at the expense of a lower immunity to changing channel conditions.

[0008] Current DMT systems allocate a fixed margin to all subchannels. For example, ADSL systems typically use a 6 dB margin on all subchannels carrying data bits. This 6 dB margin is constant on all subchannels and is independent of the type of impairment that the margin is trying to protect against.

SUMMARY OF THE INVENTION

[0009] For simplicity of reference, the systems and methods of this invention will hereinafter refer to the transceivers, or multicarrier modems, generically as modems. One such modem is typically located at a customer premises such as a home or business and is "downstream" from a central office with which it communicates. The other modem is typically located at the central office and is "upstream" from the customer premises. Consistent with industry practice, the modems are often referred to as "ATU-R" ("ADSL transceiver unit, remote," i.e., located at the customer premises) and "ATU-C" ("ADSL transceiver unit, central office," i.e., located at the central office). Each modem includes a transmitter section for transmitting data and a receiver section for receiving data, and is of the discrete multitone type, i.e., the modem transmits data over a multiplicity of subchannels of limited bandwidth. Typically, the upstream or ATU-C modem transmits data to the downstream or ATU-R modem over a first set of subchannels, which are usually the higher-frequency subchannels, and receives data from the downstream or ATU-R modem over a second, usually smaller, set of subchannels, commonly the lower-frequency subchannels.

[0010] For example, in digital subscriber line (DSL) technology, communications over a local subscriber loop between a central office and a subscriber premises is accomplished by modulating the data to be transmitted onto a multiplicity of discrete frequency carriers which are summed together and then transmitted over a subscriber loop. Individually, the carriers form discrete, non-overlapping communication subchannels which are of a limited bandwidth. Collectively, the carriers form what is effectively a broadband communications channel. At the receiver end, the carriers are demodulated and the data recovered.

[0011] DSL systems experience disturbances from other data services on adjacent phone lines, such as, for example, ADSL, HDSL, ISDN, T1, or the like. Additionally, DSL systems may experience disturbances from impulse noise, crosstalk, temperature changes, or the like. These disturbances may commence after the subject DSL service is already

initiated and, since DSL for Internet access in envisioned as a always-on service, the affects of these disturbances should be considered by the subject DSL transceiver. Additionally, the length of the phone line is a type of impairment that varies from one ADSL subscriber to another, i.e. from one ADSL installation to another, and therefore has an effect on the ADSL modem performance.

[0012] The systems and methods of this invention allow the margin in a discrete multitone modulation system to vary depending on a type of impairment. For example, this impairment can be changing over some duration or from one installation to another. Thus, different margins can be assigned to one or more of the carriers in a discrete multitone modulation communication system.

[0013] As noted above, there is a tradeoff between the robustness of the link and the achievable data rate. By setting a higher margin, a higher level of immunity to changing channel conditions is achieved at the expense of the data rate. Similarly, while a lower margin may result in a higher data rate, the immunity to changing channel conditions is reduced.

[0014] However, setting the margin equally for all subchannels at least fails to account for impairments that change over time and how the impairments may have different effects on subchannels at different frequencies. For example, temperature changes and line length effect different frequencies with differing degrees of interference.

[0015] Aspects of the present invention relate to a communications system having a plurality of margins.

[0016] Aspects of the present invention also relates to a method of assigning a plurality of margins to a communications system.

[0017] Aspects of this present invention additionally relate to multicarrier modulation systems and methods for different margins to be assigned to different subchannels to account for varying impairments.

[0018] These and other features and advantages of this invention are described in, or are apparent from, the following detailed description of the embodiments.

BRIEF DESCRIPTION OF THE DRAWINGS

[0019] The embodiments of the invention will be described in detail, with reference to the following figures wherein:

[0020] Fig. 1 is a functional block diagram illustrating an exemplary modem according to this invention; and

[0021] Fig. 2 is a flowchart outlining an exemplary method for assigning margins according to this invention.

DETAILED DESCRIPTION OF THE INVENTION

[0022] In an exemplary embodiment of the invention, the margin is set to be different on at least two subchannels in a discrete multitone modulation system. In this exemplary embodiment, subchannels which are expected to incur greater variations in impairment levels are set to have a higher margin, whereas subchannels which are expected to incur lower variations in impairment levels are set to have lower margins. As an example of this embodiment, consider an ADSL transmission system transmitting data over telephone wires and consider the case where the impairment is changing channel conditions due to temperature fluctuations. Since telephone wire is typically made out of copper, the attenuation, i.e., the insertion loss, characteristics will depend on the temperature of the wire. As the temperature of the wire increases, the attenuation, i.e., the insertion loss, will increase. Furthermore, the insertion loss also varies with frequency as the temperature changes. Therefore, as the temperature increases, in addition to an overall increase in insertion loss, the insertion loss at the higher frequencies increases more than the insertion loss at the lower frequencies. Table 1 shows a correlation of frequency versus insertion loss of an exemplary 13,500 ft. 26 AWG line at various frequencies for 70°F and 120°F.

	Frequency (kHz)										
	20	40	100	200	260	300	400	500	600	780	1100
Insertion loss (dB) at 70° F	29.8	36.7	45.2	52.8	57.3	60.2	67.7	74.8	81.7	93.0	110
Insertion loss (dB) at 120° F	31.9	39.6	49.4	57.4	61.8	64.8	72.3	79.3	86.1	97.9	116

TABLE 1: Insertion loss of 13500 ft 26 AWG line versus frequency at 70F and 120F

[0023] From Table 1, it is apparent that the difference in insertion loss from 120°F to 70°F is 2.1 dB at 20 kHz, whereas the difference in insertion loss from 120°F to 70°F is 6 dB at 1100 kHz. For this exemplary embodiment, a higher margin could be allocated to carriers at higher frequencies and a lower margin allocated to carriers at lower frequencies. For example, the carrier at 20 kHz will only need a 2.1 dB margin, because even if the temperature changes from 70°F to 120°F, the insertion loss will only change by 2.1 dB and, as a result, the system bit error rate requirement can still be met after the temperature change. Similarly, the carrier at 1100 kHz will need a 6 dB margin, since as the temperature changes from 70°F to 120°F, the insertion loss will change by 6 dB and, as a result, the system bit error rate requirement will still be satisfied even after the temperature change.

[0024] However, it is to be appreciated that the margin is not allocated to each subchannel in a fixed manner, but rather varies based on the expected change in impairments over time or as impairments vary from one DSL installation to another. However, that does not preclude the possibility that different subchannels can have the same margin assigned to them. For example, a subchannel may have a certain margin assigned based on a particular impairment, while another subchannel may have the same margin assigned based on another impairment. These impairments can include, but are not limited to, changes in the levels of crosstalk from other transmission systems, impulse noise, temperature changes, line length, radio frequency interference and other ingress, or the like. As a result, for example, since certain subchannels are not overly burdened with a common margin, the overall data rate of the system can be increased without sacrificing the robustness of the system.

[0025] For example, and with reference to Table 1, by lowering the margin of the lower carriers from 6 dB to 2.1 dB, the channel data rate has increased. This increase can occur without a loss of the immunity to temperature variations on the line since the lower frequencies are less susceptible to temperature changes than the higher frequencies. In general, the systems and methods of this invention can be adapted to set a margin for any impairment that varies over time, or is installation based, and may, for example, effect different frequencies in different ways.

[0026] As another example, consider crosstalk from another transmission system. If the crosstalking transmission system is known to use only a portion of the frequency spectrum utilized by the discrete multitone modulation system, then the margins can be decreased on the carriers that are known to be outside the frequency spectrum of the crosstalking system. For example, ISDN systems are an example of a crosstalk source for ADSL systems. ISDN systems typically transmit only up to approximately 150 kHz. Thus, for example, employing the teachings of this invention, carriers above 150 kHz can operate at lower margins than carriers below 150 kHz where the ISDN crosstalk is present.

[0027] As another example, the margin in an ADSL system can be varied depending on the length of the telephone wire. Table 2 shows a relationship of insertion loss of an exemplary 9000 ft. 26 AWG line at frequencies for 70°F and 120°F.

	Frequency (kHz)										
	20	**40**	**100**	**200**	**260**	**300**	**400**	**500**	**600**	**780**	**1100**
Insertion loss (dB) at 70° F	20.0	24.4	30.1	35.2	38.2	40.2	45.1	49.9	54.4	62.0	73.6
Insertion loss (dB) at 120° F	21.4	26.3	32.8	38.2	41.2	43.2	48.2	52.9	57.4	65.3	77.5

TABLE 2: Insertion loss of 9000 ft 26 AWG line versus frequency at 70F and 120F

[0028] Comparing Table 1 and Table 2, it is apparent that an increase in insertion loss as temperature increases depends on the length of the telephone line as well. Thus, on the exemplary 9,000 ft. phone line, a 50°F temperature change results in an average of only 2.8 dB increase in insertion loss. On the 13,500 ft. phone line, a 50°F temperature change resulted in an average of 4.3 dB increase in insertion loss. For this illustrative example, the

margin on the subchannels is varied depending on the length of the phone line. As an example, if the phone line is shorter, e.g., 9,000 ft., the average margin can be decreased on the subchannels by 4.3-2.8=1.5 dB as compared to a longer 13,500 ft. loop without sacrificing immunity to temperature changes on the phone line. This is possible because a shorter phone line will not experience as much of a change in insertion loss due to temperature changes as a longer phone line.

[0029] For this illustrative example, the margin allocated to different subchannels takes into account information about the length of the telephone line. As an example, the insertion loss difference from 70°F to 120°F at 20 kHz is 2.1 dB for the 13,500 ft. line. On the other hand, the insertion loss difference from 70°F to 120°F at 20 kHz is 1.4 dB for the 9,000 ft. line. Therefore, for this exemplary situation, a margin of 2.1 dB would be allocated to the carrier at 20 kHz on a 13,500 ft. line whereas a margin of 1.4 dB would be allocated to the carrier at 20 kHz on the 9,000 ft. line. The immunity to temperature variations on the line would be the same for both the systems operating at 9,000 ft. and 13,500 ft. As a result, the overall system data rate can be increased on shorter lines without sacrificing a loss in robustness.

[0030] Fig. 1 illustrates an exemplary embodiment of a multicarrier modem 100. In particular, the multicarrier modem 100 comprises a controller 10, a memory 20, a discrete multitone modulation system 30, a data rate determiner 40, a signal to noise ratio determiner 50, a margin determiner 60 and a margin storage 70, all interconnected by link 5. The multicarrier modem 100 is also connected to one or more computer or computer-type devices 80 and additional modems (not shown) via communications link 10. For ease of illustration, the multicarrier modem 100 has been illustrated in block diagram format with only the components needed for the exemplary embodiment of this invention. Additional information and further discussion of the operation and structure of an exemplary multicarrier modem can be found in copending U.S. Patent Application Serial No. 09/485,614 entitled "Splitterless Multicarrier Modem," incorporated herein by reference in its entirety.

[0031] While the exemplary embodiment illustrated in Fig. 1 shows the multicarrier modem 100 and various components collocated, it is to be appreciated that the various components of the multicarrier modem can be rearranged and located in whole or in part at

an ATU-R and/or ATU-C. Furthermore, it is to be appreciated, that the components of the multicarrier modem 100 can be located at various locations within a distributed network, such as a POTS network, or other comparable telecommunications network. Thus, it should be appreciated, that the components of the multicarrier modem 100 can be combined into one device or distributed amongst a plurality of devices. As will be appreciated from the following description, and for reasons of computational efficiency, the components of the multicarrier modem can be arranged at any location within a telecommunications network and/or modem without affecting the operation of the system.

[0032] The links 5 and 10 can be a wired or a wireless link or any other known or later developed element(s) that is capable of supplying and communicating electronic data to and from the connected elements. Additionally, the computer device 80, can be, for example, a personal computer or other device. In general, the computer device 80 can be any device that uses a modem to transmit and/or receive data.

[0033] In operation, the multicarrier modem 100 is installed, for example, in a customer premises or in a central office. During this installation, certain fixed quantities such as line length are known and can be stored in the multicarrier modem 100. During an initial installation, or at any subsequent time for which a redetermination in margins is appropriate, for example, based on an increased bit error rate, changes in the signal to noise ratio, seasonal changes, or the like, the controller 10, in cooperation with the memory 20, the discrete multitone modulation system 30 and the margin determiner 60 can determine and store margins. For example, as illustrated above in exemplary Tables 1 and 2, margins can be determined for temperature fluctuations and the length of the wire line based on, for example, the actual installation and historical data. Furthermore, routines can be established by the margin determiner 60 to evaluate and compile statistical information relating to one or more carriers. For example, this statistical information can be compiled during modem idle times in response to impairments seen on the one or more carriers. This statistical information can then be used to determine appropriate margins for one or more carriers.

[0034] Alternatively the modem may measure the noise on the line during idle times and determine that a particular type of crosstalker, e.g., another ADSL or HDSL modem, is present. Since the spectral content of these types of crosstalkers are known, this information can be used to determine the margin. For example, if the crosstalker is an ATU-R ADSL

modem then it is known that ATU-R ADSL modems transmit approximately in the 20-130 kHz range. This information can be used to determine the margin for the carriers in the 20-130 kHz frequency range.

[0035] Alternatively, a predetermined set of margins, for example, for known impairments, can be downloaded from, for example, a central office modem or other location within a communications network. The determined and/or downloaded margins are then stored in the margin storage 70. Similarly, groups of margins can be stored based on, for example, geographic information, seasonal information, line length information, or the like.

[0036] During training of the multicarrier modem 100, the SNR determiner 50, in cooperation with the controller 10, the memory 20, and the DMT system 30, determines the signal to noise ratio of the carriers. Knowing the signal to noise ratio of the carriers, the data rate determiner 40 determines the raw data rate of the carriers based on the signal to noise ratio and the bit error rate. This raw data rate reflects the data rate of carriers with no margin.

[0037] Generally, the bit error rate is set in advance, for example, by the manufacturer. Additionally, the data rate is generally governed by a range that is, for example, guaranteed as a maximum, by a DSL provider. Therefore, based on the set bit error rate, the signal to noise ratio for a known quantity of bits can be determined.

[0038] Knowing the signal to noise ratio, the margins for the carriers can be set, for example, based on one or more, or a combination of, entered criteria or determined criteria. For example, an entered criteria can be based on the loop length. A determined criteria can be, for example, based on standard temperature variance information that can, for example, be downloaded from the service provider. Alternatively, for example, the margins can be set based on historical data that relates to, for example, impairments on the line. In general, the margins can be set such that a balance between the data rate and the impairment immunity is maximized.

[0039] Having retrieved the margins for one or more of the carriers, the margins are set in the DMT system 30. The margins can then be subtracted from the carrier to determine an

updated data rate for each carrier. Having set the margins, and knowing the data rate, the DMT system can then commence communication over the communications link 10.

[0040] Fig. 2 illustrates an exemplary method of assigning margins to carriers according to an exemplary embodiment of this invention. In particular, control begins in step S100 and continues to step S110. In step S110, a determination is made whether margins are to be determined. If margins are to be determined, control continues to step S120. Otherwise, control jumps to step S130.

[0041] In step S120, the margins are determined and stored. Control then continues to step S130.

[0042] In step S130, the signal to noise ratio of the carriers are determined. Next, in step S140, the raw data rate of the carriers is determined based on the signal to noise ratio and the bit error rate. Next, in step S150, the margins for the carriers are retrieved. Control then continues to step S160.

[0043] In step S160, the margins for the carriers are set. Next, in step S170, the margins are subtracted from the carriers to determine an updated data rate for each carrier. Control then continues to step S180.

[0044] In step S180, communications commence. Control then continues to step S190 where the control sequence ends.

[0045] However, it is to be appreciated that the steps in Fig. 2 need not occur in the order illustrated. For example, at any point in time there could be an option to re-determine the margins. Similarly, based on, for example, the time of the day, day, location, error rate, service provider directive, a change in the quality of service requirement, or the like, the margins could be adjusted. Alternatively, at any time, updated margins could be downloaded and stored in the margin storage. Alternatively, if it known that margins will be incorporated in the determination of the data rate, step S140 could be bypassed since it is known that the raw data rate will not be used.

[0046] Furthermore, the systems and methods of this invention can also apply to any multicarrier modulation based communication system including wireless LANs, such as wireless LAN 802.11 and ETSI Hyperlan standards, wireless access systems, home and access power-line communication systems, or the like.

[0047] As illustrated in Fig. 1, the multicarrier modem and related components can be implemented either on a DSL modem, or a separate program general purpose computer having a communications device. However, the multicarrier modem can also be implemented in a special purpose computer, a programmed microprocessor or microcontroller and peripheral integrated circuit element, and ASIC or other integrated circuit, a digital signal processor, a hardwired or electronic or logic circuit such as a discrete element circuit, a programmable logic device such as a PLD, PLA, FPGA, PAL, or the like, and associated communications equipment. In general, any device capable of implementing a finite state machine that is in turn capable of implementing the flowchart illustrated in Fig. 2 can be used to implement the multicarrier modem 100 according to this invention.

[0048] Furthermore, a disclosed method may be readily implemented in software using object or object-oriented software development environment that provides portable source code that can be used on a variety of computers, workstations, or modem hardware platforms. Alternatively, the disclosed modem may be implemented partially or fully in hardware using standard logic circuits or a VLSI design. Other software or hardware can be used to implement the systems in accordance with this invention depending on the speed and/or efficiency requirements of the systems, the particular function, and the particular software or hardware systems or microprocessor or microcomputer systems being utilized. The multicarrier modem illustrated herein, however, can be readily implemented in hardware and/or software using any known or later developed systems or structures, devices and/or software by those of ordinary skill in the applicable art from the functional description provided herein and with a general basic knowledge of the computer and telecommunications arts.

[0049] Moreover, the disclosed methods can be readily implemented as software executed on a programmed general purpose computer, a special purpose computer, a microprocessor and associated communications equipment, or the like. In these instances, the methods and systems of this invention can be implemented as a program embedded on a modem, such as a DSL modem, or the like. The multicarrier modem can also be implemented by physically incorporating the system and method in a software and/or hardware system, such as a hardware and software system of a modem, such as an ADSL modem, or the like.

[0050] It is, therefore, apparent that there has been provided in accordance with the present invention, systems and methods for assigning margins to carriers. While this invention has been described in conjunction with a number of embodiments, it is evident that many alternatives, modifications and variations would be or are apparent to those of ordinary skill in the applicable art. Accordingly, Applicants intend to embrace all such alternatives, modifications, equivalents and variations that are within the spirit and the scope of this invention.

What is Claimed is:

1. A multicarrier modulation communication system comprising:

a plurality of subchannels; and

a plurality of margins.

2. The system of claim 1, wherein the plurality of margins are based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, wire line length, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

3. The system of claim 1, wherein the plurality of margins are at least one of an average margin and a subchannel specific margin.

4. The system of claim 3, wherein the average margin is applied equally to a portion of the plurality of subchannels.

5. The system of claim 1, further comprising a margin determiner that determines at least one margin.

6. The system of claim 1, further comprising a margin storage device that stores at least one margin.

7. A multicarrier modulation communication system comprising:

a plurality of subchannels; and

a plurality of margins, wherein one of the plurality of margins is assigned to at least one of the plurality of subchannels.

8. The system of claim 7, wherein the plurality of margins are based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, wire line length, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

9. The system of claim 7, wherein the plurality of margins are at least one of an average margin and a subchannel specific margin.

10. The system of claim 9, wherein the average margin is applied equally to a portion of the plurality of subchannels.

11. The system of claim 7, further comprising a margin determiner that determines at least one margin.

12. The system of claim 7, further comprising a margin storage device that stores at least one margin.

13. A multicarrier modulation communication system communicating on a wire line over a plurality of subchannels, wherein at least one margin based on a length of the wire line is assigned to at least one of the plurality of subchannels.

14. The system of claim 13, wherein the at least one margin is based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

15. The system of claim 13, wherein the at least one margin is based on at least one of an average margin and a subchannel specific margin.

16. The system of claim 15, wherein the average margin is applied equally to a portion of the plurality of subchannels.

17. The system of claim 13, further comprising a margin determiner that determines at least one margin.

18. The system of claim 13, further comprising a margin storage device that stores at least one margin.

19. A multicarrier modulation communication system having a plurality of subchannels, wherein at least two subchannels have a different margin.

20. The system of claim 19, wherein the margin is based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, wire line length, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

21. The system of claim 19, wherein the margins are at least one of an average margin and a subchannel specific margin.

22. The system of claim 21, wherein the average margin is applied equally to a portion of the at least two subchannels.

23. An information storage media comprising margin information for a mulitcarrier modulation system having a plurality of subchannels, wherein at least two subchannels have a different margin.

24. A method of enhancing multicarrier modulation communication over a plurality of subchannels comprising communicating over the plurality of subchannels using at least two different margins.

25. The method of claim 24, wherein the at least two margins are based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, wire line length, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

26. The method of claim 24, wherein the at least two margins are at least one of an average margin and a subchannel specific margin.

27. The method of claim 26, wherein the average margin is applied equally to a portion of the plurality of subchannels.

28. The method of claim 24, further comprising determining at least one margin.

29. The method of claim 24, further comprising storing at least one margin.

30. A method for multicarrier modulation communication over a plurality of subchannels comprising:

selecting a first number of the subchannels;

assigning a first margin to the first number of the subchannels;

selecting a second number of the subchannels; and

assigning a second margin to the second number of subchannels, wherein the first margin and the second margin are different.

31. The method of claim 30, wherein the margins are based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, wire line length, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

32. The method of claim 30, wherein the margins are at least one of an average margin and a subchannel specific margin.

33. The method of claim 32 wherein the average margin is applied equally to a portion of either the first or second number of subchannels.

34. The method of claim 30, further comprising a margin determiner that determines at least one margin.

35. The method of claim 30, further comprising a margin storage device that stores at least one margin.

36. A method for multicarrier modulation communication over a wire line using a plurality of subchannels, wherein at least one margin based on a length of the wire line is assigned to at least one of the plurality of subchannels.

37. The method of claim 36, wherein the at least one margin is based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

38. The method of claim 36, wherein the at least one margin is at least one of an average margin and a subchannel specific margin.

39. The method of claim 38, wherein the average margin is applied equally to a portion of the plurality of subchannels.

40. The method of claim 36, further comprising a margin determiner that determines at least one margin.

41. The method of claim 36, further comprising a margin storage device that stores at least one margin.

42. A method for communicating in a multicarrier modulation communications environment having at least two subchannels, wherein at least two of the at least two subchannels have a different margin.

43. The method of claim 42, wherein the margins are based on at least one of changes in the levels of a crosstalk, impulse noise, temperature changes, wire line length, radio frequency interference, a bit error rate, a signal to noise ratio, a seasonal change, statistical information, time information, day information and data rate information.

44 The method of claim 42, wherein the margins are at least one of an average margin and a subchannel specific margin.

45. The method of claim 42, wherein the average margin is applied equally to a portion of the at least two subchannels.

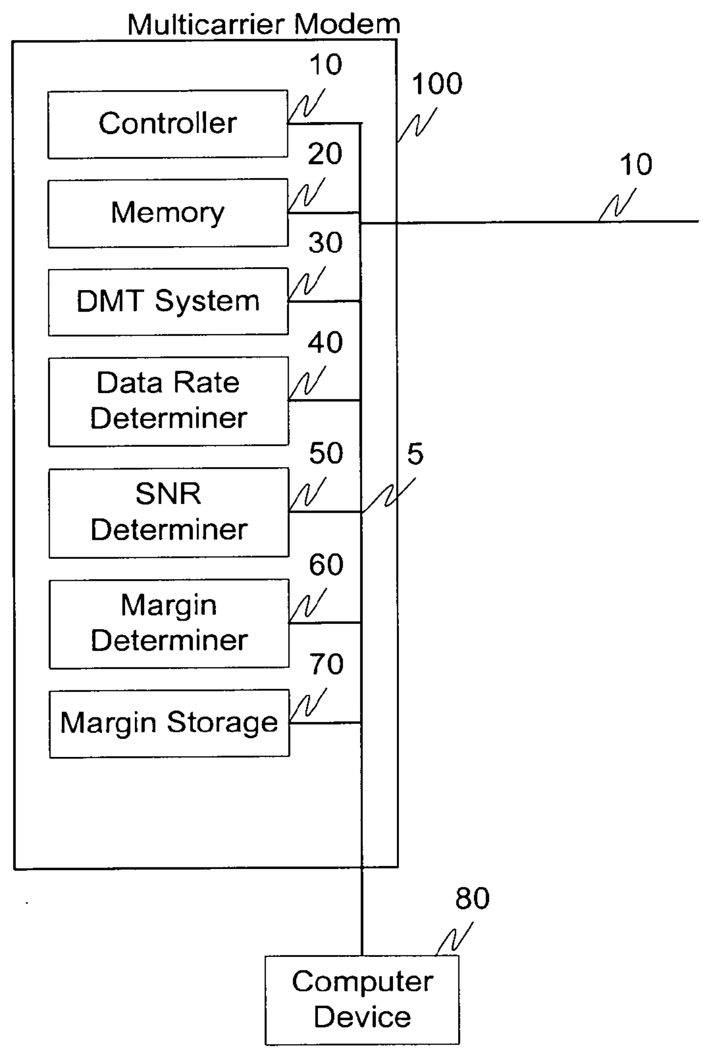

Multicarrier Modem
10
100
Controller
20
10
Memory
30
DMT System
40
Data Rate
Determiner
50
5
SNR
Determiner
60
Margin
Determiner
70
Margin Storage
80
Computer
Device

Fig. 1

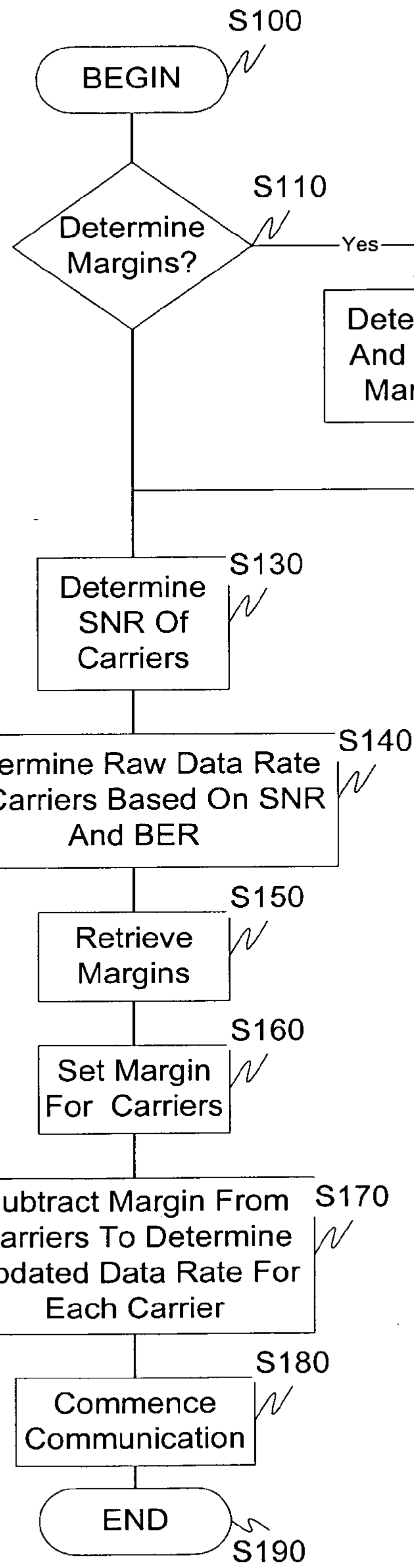

S100
BEGIN
S110
Determine Margins?
Yes
S120
Determine And Store Margins
S130
Determine SNR Of Carriers
S140
Determine Raw Data Rate Of Carriers Based On SNR And BER
S150
Retrieve Margins
S160
Set Margin For Carriers
S170
Subtract Margin From Carriers To Determine Updated Data Rate For Each Carrier
S180
Commence Communication
END
S190

Fig. 2

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 01/12555

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L27/26

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, PAJ, WPI Data, INSPEC, COMPENDEX

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	WO 99 20027 A (AWARE) 22 April 1999 (1999-04-22) page 11, line 18 - line 23 page 11, line 26 ---	19-29
X	COLIN ET AL.: "QoS considerations for DMT-based ADSL and VDSL systems" IEEE INTERNATIONAL CONFERENCE ON ACOUSTICS, SPEECH AND SIGNAL PROCESSING, 12 - 15 May 1998, pages 3437-3440, XP000951196 New York, US ISBN: 0-7803-4429-4 page 3438, left-hand column, paragraph 4 ---	19-29
X	EP 0 918 422 A (MOTOROLA INC) 26 May 1999 (1999-05-26) column 4, line 6 - line 17 --- -/--	19-29

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance
"E" earlier document but published on or after the international filing date
"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)
"O" document referring to an oral disclosure, use, exhibition or other means
"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.
"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
26 September 2001	09/10/2001
Name and mailing address of the ISA European Patent Office, P.B. 5818 Patentlaan 2 NL - 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Authorized officer Scriven, P

1

INTERNATIONAL SEARCH REPORT

ational Application No
rul/US 01/12555

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
X	US 5 822 374 A (LEVIN) 13 October 1998 (1998-10-13) figure 3 ---	19-29
X	EP 0 955 744 A (NEC) 10 November 1999 (1999-11-10) page 25, line 30 - line 31 page 26, line 15 - line 19 ---	19-29
A	WO 98 57472 A (AWARE) 17 December 1998 (1998-12-17) page 5, line 10 - line 11 ---	19,23,24
A	"Spectrum management for loop transmission systems" ANSI DRAFT , ‘Online! 1999, pages i-110, XP002178517 Retrieved from the Internet: <URL:http://ftp.tiaonline.org/tr-41/TR419/WORKING/8-99/-046%20T1E1.pdf> ‘retrieved on 2001-09-26! paragraph ‘4.3.5.1! -----	20,25

FURTHER INFORMATION CONTINUED FROM PCT/ISA/ 210

Continuation of Box I.2

Claims Nos.: 1-18, 30-45

In view of the large number of independent claims, and also of their wording, which renders it difficult, if not impossible, to determine the matter for which protection is sought, the present application fails to comply with the clarity and conciseness requirements of Article 6 PCT (see also Rule 6.1(a) PCT) to such an extent that a meaningful search is impossible. Consequently, the search has been carried out for those parts of the application which do appear to be clear and concise, namely claims 19-29. These claims define the use of at least two different margins. This feature is essential, as is specified in paratraph '0012! of the description.

The applicant's attention is drawn to the fact that claims, or parts of claims, relating to inventions in respect of which no international search report has been established need not be the subject of an international preliminary examination (Rule 66.1(e) PCT). The applicant is advised that the EPO policy when acting as an International Preliminary Examining Authority is normally not to carry out a preliminary examination on matter which has not been searched. This is the case irrespective of whether or not the claims are amended following receipt of the search report or during any Chapter II procedure.

INTERNATIONAL SEARCH REPORT

Information on patent family members

ational Application No
PCT/US 01/12555

Patent document cited in search report		Publication date	Patent family member(s)		Publication date
WO 9920027	A	22-04-1999	AU	1077699 A	03-05-1999
			EP	1021901 A2	26-07-2000
			WO	9920027 A2	22-04-1999
EP 0918422	A	26-05-1999	US	6122247 A	19-09-2000
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			AU	7827398 A	30-12-1998
			EP	1013043 A1	28-06-2000
			WO	9857472 A1	17-12-1998

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
10 January 2002 (10.01.2002)

PCT

(10) International Publication Number
WO 02/03557 A1

(51) International Patent Classification[7]: H04B 1/38

(21) International Application Number: PCT/US01/11309

(22) International Filing Date: 5 April 2001 (05.04.2001)

(25) Filing Language: English

(26) Publication Language: English

(30) Priority Data:
09/609,591 30 June 2000 (30.06.2000) US

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(81) Designated States (*national*): AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NO, NZ, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TR, TT, TZ, UA, UG, UZ, VN, YU, ZA, ZW.

(84) Designated States (*regional*): ARIPO patent (GH, GM, KE, LS, MW, MZ, SD, SL, SZ, TZ, UG, ZW), Eurasian patent (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European patent (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE, TR), OAPI patent (BF, BJ, CF, CG, CI, CM, GA, GN, GW, ML, MR, NE, SN, TD, TG).

Published:
— *with international search report*

For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

WO 02/03557 A1

(54) **Title:** METHOD AND SYSTEM FOR MODE ADAPTATION IN WIRELESS COMMUNICATION

(57) **Abstract:** A method and system for selective mode adaptation for transmitting data by spatial multiplexing applicable in communications systems with a transmit unit (12) having multiple transmit antennas (18) or multiple transmit units (204,206) and a receive unit (14) having multiple receive antennas (34). A channel descriptor (22), such as channel matrix H or a channel matrix filter H, with has sub-descriptors corresponding to the transmit antennas (18) is determined and a quality parameter, such as signal-to-interference and noise ratio, signal-to-noise ratio or power level are chosen. The quality parameter is assigned a threshold and the sub-descriptor or sub-descriptors whose quality parameters do not meet the threshold are identified and deactivated.

METHOD AND SYSTEM FOR MODE ADAPTATION IN WIRELESS COMMUNICATION

FIELD OF THE INVENTION

The present invention relates generally to wireless communication systems and methods, and more particularly to mode adaptation including selection of transmit antennas in transmit units employing multiple antennas for spatial multiplexing.

BACKGROUND OF THE INVENTION

Wireless communication systems serving stationary and mobile wireless subscribers are rapidly gaining popularity. Numerous system layouts and communications protocols have been developed to provide coverage in such wireless communication systems.

Wireless communications channels between transmit and receive devices are inherently variable and their quality fluctuates. Specifically, the quality parameters of such communications channels vary in time. Under good conditions wireless channels exhibit good communication parameters, e.g., large data capacity, high signal quality, high spectral efficiency and throughput. At these times significant amounts of data can be transmitted via the channel reliably. However, as the channel changes in time, the communication parameters also change. Under altered conditions former data rates, coding techniques and data formats may no longer be feasible. For example, when channel performance is degraded the transmitted data may experience excessive corruption yielding unacceptable

communication parameters. For instance, transmitted data can exhibit excessive bit-error rates or packet error rates. The degradation of the channel can be due to a multitude of factors such as general noise in the channel, multi-path fading, loss of line-of-sight path, excessive Co-Channel Interference (CCI) and other factors.

By reducing CCI the carrier-to-interference (C/I) ratio can be improved and the spectral efficiency increased. Specifically, improved C/I ratio yields higher per link bit rates, enables more aggressive frequency re-use structures and increases the coverage of the system.

It is also known in the communication art that transmit units and receive units equipped with antenna arrays, rather than single antennas, can improve receiver performance. Antenna arrays can both reduce multipath fading of the desired signal and suppress interfering signals or CCI. Such arrays can consequently increase both the range and capacity of wireless systems. This is true for wireless cellular telephone and other mobile systems as well as Fixed Wireless Access (FWA) systems.

In mobile systems, a variety of factors cause signal degradation and corruption. These include interference from other cellular users within or near a given cell. Another source of signal degradation is multipath fading, in which the received amplitude and phase of a signal varies over time. The fading rate can reach as much as 200 Hz for a mobile user traveling at 60 mph at PCS frequencies of about 1.9 GHz. In such environments, the problem is to

cleanly extract the signal of the user being tracked from the collection of received noise, CCI, and desired signal portions summed at the antennas of the array.

In FWA systems, e.g., where the receiver remains stationary, signal fading rate is less than in mobile systems. In this case, the channel coherence time or the time during which the channel estimate remains stable is longer since the receiver does not move. Still, over time, channel coherence will be lost in FWA systems as well.

Antenna arrays enable the system designer to increase the total received signal power, which makes the extraction of the desired signal easier. Signal recovery techniques using adaptive antenna arrays are described in detail, e.g., in the handbook of Theodore S. Rappaport, *Smart Antennas, Adaptive Arrays, Algorithms, & Wireless Position Location;* and Paulraj, A.J. et al., "Space-Time Processing for Wireless Communications", IEEE Signal Processing Magazine, Nov. 1997, pp. 49-83.

Prior art wireless systems have employed adaptive modulation of the transmitted signals with the use of feedback from the receiver as well as adaptive coding and receiver feedback to adapt data transmission to changing channel conditions. However, effective maximization of channel capacity with multiple transmit and receive antennas is not possible only with adaptive modulation and/or coding.

In U.S. Pat. Nos. 5,592,490 to Barratt et al., 5,828,658 to Ottersten et al., and 5,642,353 Roy III, teach about spectrally efficient high capacity wireless communication systems using multiple antennas at the transmitter; here a Base Transceiver Station (BTS) for Space Division Multiple Access (SDMA). In these systems the users or receive units have to be sufficiently separated in space and the BTS uses its transmit antennas to form a beam directed towards each receive unit. The transmitter needs to know the channel state information such as "spatial signatures" prior to transmission in order to form the beams correctly. In this case spatial multiplexing means that data streams are transmitted simultaneously to multiple users who are sufficiently spatially separated.

The disadvantage of the beam-forming method taught by Barratt et al., Ottersten et al., and Roy III is that the users have to be spatially well separated and that their spatial signatures have to be known. Also, the channel information has to be available to the transmit unit ahead of time and the varying channel conditions are not effectively taken into account. Finally, the beams formed transmit only one stream of data to each user and thus do not take full advantage of times when a particular channel may exhibit very good communication parameters and have a higher data capacity for transmitting more data or better signal-to-noise ratio enabling transmission of data formatted with a less robust coding scheme.

U.S. Pat. No. 5,687,194 to Paneth et al. describes a Time Division Multiple Access (TDMA) communication system using

multiple antennas for diversity. The proposed system exploits the concept of adaptive transmit power and modulation. The power and modulation levels are selected according to a signal quality indicator fed back to the transmitter.

Addressing the same problems as Paneth et al., U.S. Pat. No. 5,914,946 to Avidor et al. teaches a system with adaptive antenna beams. The beams are adjusted dynamically as the channel changes. Specifically, the beams are adjusted as a function of a received signal indicator in order to maximize signal quality and reduce the system interference.

The prior art also teaches using multiple antennas to improve reception by transmitting the same information, i.e., the same data stream from all antennas. Alternatively, the prior art also teaches that transmission capacity can be increased by transmitting a different data stream from each antenna. For more information about capacity increases achievable by transmitting different data streams from different antennas the reader is referred to U.S. Pat. No. 5,345,599 to Paulraj, A.J. et al., and to Foschini, G.J., "Layered Space-Time Architecture for Wireless Communication in a Fading Environment when Using Multi-Element Antennas", Bell Labs Technical Journal, Autumn 1996. These two approaches are commonly referred to as antenna diversity schemes and spatial multiplexing schemes.

Adaptive modulation and/or coding in multiple antenna systems involve mapping of data converted into appropriate symbols to the antennas of the transmit antenna array for transmission. In the case of spatial multiplexing there can be no coordination between transmitters, since the transmit antennas can belong to separate base stations or mobiles rather than to a single antenna array. Prior art systems do not teach rules suitable for determining antenna mappings, or, more precisely, antenna selection for varying channel conditions in the spatial multiplexing mode. Development of rules for selecting antennas in the spatial multiplexing mode would represent a significant advance in the art.

SUMMARY

The present invention provides a method for selecting antennas for transmitting data over a channel by employing spatial multiplexing, i.e., by transmitting different data streams from different antennas. The method is applicable in communications systems with a transmit unit having multiple transmit antennas or multiple transmit units, e.g., multiple transceiver stations, and a receive unit having multiple receive antennas. The first step of the method involves determining a channel descriptor. In one embodiment the descriptor is a channel matrix H. In another embodiment, where the channel is subject to time delay spread, the descriptor is a channel matrix filter $\mathbf{H}$. The channel descriptor has sub-descriptors corresponding to the transmit antennas. Specifically, channel matrix H has sub-descriptors in the form of sub-matrices h_i corresponding

to the transmit antennas. Channel matrix filter H has sub-descriptors in the form of sub-matrix filters h_i also corresponding to the transmit antennas.

The method then calls for choosing a quality parameter and a threshold for the quality parameter. In the next steps, a sub-descriptor which does not meet the threshold is identified and the antenna from among the transmit antennas associated with the sub-descriptor is deactivated.

The quality parameter used in the method can be selected from among signal-to-interference and noise ratio, signal-to-noise ratio and power level. The threshold is typically a minimum acceptable value of the quality parameter. This threshold can be re-set or adjusted and the steps of identifying another sub-descriptor or set of sub-descriptors not meeting the threshold can be repeated. The corresponding transmit antennas are then deactivated.

In case the spatial multiplexed communication is of the type employing a number of sub-carrier tones rather than just one carrier frequency, the sub-descriptors are associated with an average value of the quality parameter. In particular, the average value is the average of the quality parameter over the sub-carrier tones. It is then this average value of the quality parameter which is compared with the threshold to determined whether the threshold is met.

In one embodiment of the method, the sub-matrix h_j is removed from the channel matrix H to obtain a subset channel matrix H'. The remaining sub-matrices h_i can be rearranged after removal of h_j. In particular, the remaining sub-matrices h can be ordered in accordance with the threshold, i.e., in descending order starting with the sub-matrix h_i which exceeds the threshold the most or has the best quality parameter.

The data transmitted is typically coded and modulated in accordance with a selected mode. The mode is characterized by a coding rate and a modulation. The setting of the threshold can be based on the selected mode and the selected mode can be based on the quality parameter.

The receive unit can employ any receiver such as a maximum likelihood receiver, a zero forcing equalizer receiver, a successive cancellation receiver, a minimum mean square error equalizer (MMSE) receiver.

In another embodiment, a set of sub-descriptors, i.e., a set of sub-matrices h_i or set of sub-matrix filters h_i is identified and transmit unit antennas associated with that set are all deactivated at one time, rather than one by one. This method can be applied in steps as well. A set of antennas or a single antenna not meeting the assigned threshold can be deactivated in each step.

In still another embodiment, the transmit antennas belong to separate transmit units, e.g., to different base

stations. In this case the sub-matrix h_j can represent the base station transceiver which is to be deactivated to improve the quality parameter at the receive unit.

Systems of the invention can be used for spatial multiplexed communications between transmit units with multiple antennas and receive units with multiple antennas, or between separate transceivers, e.g., base stations, and receive units with multiple antennas. The receive unit has a channel estimation block for determining channel descriptor, a mode selection block for receiving the quality parameter, assigning a threshold to the quality parameter, and identifying among the sub-descriptors of the channel descriptor at least one sub-descriptor not meeting the threshold. The receive unit has a feedback unit for sending feedback related to the at least one sub-descriptor to the transmit unit or, as the case may be, to the base stations.

In case the transmit unit has an antenna array, a controller at the transmit unit receives the feedback and deactivates the corresponding transmit antennas. In case the transmit antennas belong to separate transceivers, a common controller or separate control units can be used to receive the feedback and deactivate the corresponding transceivers or their antennas.

The method of the invention can be employed in multi-tone communications using a number of sub-carrier tones for transmitting data from each transmit antenna.

A detailed description of the invention and the preferred and alternative embodiments is presented below in reference to the attached drawing figures.

BRIEF DESCRIPTION OF THE FIGURES

DETAILED DESCRIPTION

The method and wireless systems of the invention will be best understood after first considering the high-level diagrams of Figs. 1 and 2. Fig. 1 illustrates a portion of a wireless communication system **10**, e.g., a cellular

wireless system. For explanation purposes, the downlink communication will be considered where a transmit unit **12** is a Base Transceiver Station (BTS) and a receive unit **14** is a mobile or stationary wireless user device. Exemplary user devices include mobile receive units **14A, 14B, 14C** which are portable telephones and car phones and a stationary receive unit **14D,** which can be a wireless modem unit used at a residence or any other fixed wireless unit. Of course, the same method can be used in uplink communication from wireless units **14** to BTS **12.**

BTS **12** has an antenna array **16** consisting of a number of transmit antennas **18A, 18B, ..., 18M.** Receive units **14** are equipped with antenna arrays **20** of N receive antennas (for details see Figs. 2, 3 and 5). BTS **12** sends transmit signals TS to all receive units **14** via channels **22A** and **22B.** For simplicity, only channels **22A, 22B** between BTS **12** and receive units **14A, 14B** are indicated, although BTS **12** transmits TS signals to all units shown. In this particular case receive units **14A, 14B** are both located within one cell **24.** However, under suitable channel conditions BTS **12** can transmit TS signals to units outside cell **24.**

The time variation of channels **22A, 22B** causes transmitted signals TS to experience fluctuating levels of attenuation, interference, multi-path fading and other deleterious effects. Therefore, communication parameters of channels **22A, 22B** such as data capacity, signal quality, spectral efficiency or throughput undergo temporal changes. Thus, channels **22A, 22B** can not at all times support efficient

propagation of high data rate signals TS or signals which are not formatted with a robust coding algorithm.

In accordance with the invention, antenna array **16** at BTS **12** employs spatial multiplexing, reduces interference, increases array gain and achieves other advantageous effects. Antenna arrays **20** at receive units **14** are set up to receive the spatial multiplexed signals from BTS **12.** The method of the invention finds an optimum choice of transmit antennas **18A, 18B, ..., 18M** selected adaptively with changing conditions of channels **22A, 22B.** In other words, the method of the invention implements an adaptive and optimal selection of transmit antennas **18A, 18B, ...,** **18M,** deactivating some of these antennas in accordance with the rules described below to improve performance.

Fig. 2 illustrates the fundamental blocks of transmit unit **12** and one receive unit **14** necessary to employ the method. Transmit unit **12** has a control unit **26** connected to a data processing block **28** for receiving data **30** to be converted to spatially multiplexed transmit signals TS to select transmit antennas **18A, 18B, ..., 18M** for transmission therefrom. An up-conversion and RF amplification block **32** supplies the transmit signals TS to antennas **18A, 18B, ...,** **18M.**

On the other side of the link, receive unit **14** has N receive antennas **34A, 34B, ..., 34N** in its array **20** for receiving receive signals RS. An RF amplification and down-conversion block **36** processes receive signals RS and passes them to data processing block **38.** Data processing

block **38** includes a channel measurement or estimation unit (see Fig. 5) which obtains a measurement of the channel coefficients matrix H characterizing channel **22.**

A mode selection block **40** uses matrix H and a chosen quality of service QoS or quality parameter QP to determine which of transmit antennas **18** should be deactivated to improve reception. The quality parameter QP used by block **40** can be any useful signal characteristics measure such as signal-to-interference and noise ratio (SINR), signal-to-noise ratio (SNR), power level. Block **40** makes the determination about which of transmit antennas **18A, 18B,** ... **18M** should be transmitting in order to keep the quality parameter above a certain minimum required value or threshold. This selection is fed back as indicated by dashed line **42** to transmit unit **12.** In case channel **22** is a time-division duplexed (TDD) channel, which is reciprocal between the receive and transmit units, no separate feedback is required. In response, unit **26** switches off or deactivates the transmit antennas which block **40** has determined should be deactivated.

An exemplary embodiment of a transmit unit **50** for practicing the method of the invention is shown in Fig. 3. Data **52,** in this case in the form of a binary stream, has to be transmitted. Before transmission, data **52** may be interleaved and pre-coded by interleaver and pre-coder (not shown). The purpose of interleaving and pre-coding is to render the data more robust against errors. Both of these techniques are well-known in the art.

Data **52** is delivered to a conversion unit, more specifically a spatial multiplexing block **56**. Block **56** converts data **52** into k streams of symbols at chosen modulation rates and coding rates. For example, data **52** can be converted into symbols through modulation in a constellation selected from among PSK, QAM, GMSK, FSK, PAM, PPM, CAP, CPM or other suitable constellations. The transmission rate or throughput of data **52** will vary depending on the modulation and coding rates used in each of the k streams.

Mode	Modulation Rate (bits/symbol)	Coding Rate	Throughput (bits/s/Hz)
1	2	3/4	3/2
2	2	2/3	4/3
3	2	1/2	1
4	2	1/3	2/3
5	4	3/4	3
6	4	2/3	8/3
7	4	1/2	2
8	4	1/3	4/3
9	5	3/4	15/4
10	5	2/3	10/3
11	5	1/2	5/2
12	5	1/3	5/3
13	6	3/4	9/2
14	6	2/3	4
15	6	1/2	3
16	6	1/3	2

Table 1 illustrates some typical modulation and coding rates with the corresponding throughputs which can be used in the spatial multiplexing method of the invention. The entries are conveniently indexed by a mode number.

The mode column can be used to more conveniently identify the modulation and coding rates which are to be applied to the k streams. Tables analogous to Table 1 for other coding rates and modulation can be easily derived as these techniques are well-known in the art.

Once coded and modulated in symbols, data **52** passes to a switching unit **60**. Depending on its setting, switching unit **60** routs modulated and coded k streams of spatially multiplexed data **52** to all or a subset of its M outputs. The M outputs lead to the corresponding M transmit antennas **72** via an up-conversion and RF amplification stage **70** having individual digital-to-analog converters and up-conversion/RF amplification blocks **74**. Transmit antennas **72** transmit data **52** in the form of transmit signals TS. In this case transmit antennas T_1, T_2, ... T_M with the exception of transmit antenna T_j are transmitting coded streams. In other words, k=M-1. The determination to deactivate antenna T_j is made in accordance with the method of the invention as described below.

Transmit unit **50** also has a controller **66** connected to spatial multiplexing block **56** and to switching unit **68**. A database **78** is connected to controller **66**. Database **78** conveniently contains a table, e.g., a spatial multiplexing look-up table indexed by mode as in exemplary table 1. The convenience of indexing by mode resides in the fact that feedback to transmit unit **50** does not require much bandwidth.

Specifically, transmit unit **50** receives feedback from receive unit **90** (see Fig. 5) via a feedback extractor **80**. Feedback extractor **80** detects an antenna number or any other designation which antennas to operate and which to deactivate and forwards this information to controller **66**. In some embodiments feedback extractor detects mode number and associated antenna number, and forwards it to controller **66**. Controller **66** looks up the mode number in database **78** and thus determines the modulation, coding rate and any other parameters for the associated antenna.

Receive unit **90** can send back a channel descriptor, e.g., a channel matrix H, a channel matrix filter H or some other suitable descriptor identifying the action of the channel on transmitted signals TS, to transmit unit **50**. In these cases transmit unit **50** can use the channel descriptor in its operation to derive any information in addition to antenna number and mode number to adapt its transmission to channel **22**. In the event of using time-division duplexing (TDD), the feedback information, i.e., the channel parameters are obtained during the reverse transmission from receive unit **90** or remote subscriber unit, as is known in the art, and no dedicated feedback extractor **80** is required.

When channel **22** experiences delay spread, it can be modeled as a Finite Impluse Response (FIR) channel, i.e., channel **22** has a memory and any representation of channel **22** should have a time dimension. Depending on the transmission symbol rate, for a given delay spread channel **22** will have

some number L of symbol delay taps. When there is no delay spread channel **22** can be represented by an NxM matrix where N is the number of receive antennas **92** (see Fig. 5) and M is the number of transmit antennas **72**. When there is delay spread, channel **22** can be represented by a matrix filter **H** which is constructed of H_1, H_2, ... H_L, where H_i is the NxM channel matrix at i-th delay tap.

When multi-carrier modulation such as OFDM is used, the symbol duration is chosen much longer than the channel delay spread. In this case, each sub-carrier frequency or tone has an individual channel represented by an NxM matrix, i.e., H_1, H_2, ... H_T where T is the number of sub-carrier tones.

In an embodiment of the invention where inter-symbol interference (ISI) is not a problem, the parameters of channel **22** are expressed by a single channel matrix H. In accordance with this descriptor, transmit signals TS propagating through channel **22** are affected by channel coefficients a_{xy} of matrix H. Matrix H is composed of sub-matrices h, here in the form of columns labeled h_1, h_2, ..., h_j, ... h_M. Each antenna T_1, T_2, ..., T_j, ..., T_M is associated with a corresponding sub-matrix h_1, h_2, ..., h_j, ... h_M. The dimension or number of entries in each sub-matrix h is dictated by the number of receive antennas **92** employed by receive unit **90** (see Fig. 5); in this case the number is N. Hence, channel coefficients matrix H is an NxM matrix with channel coefficients a_{xy} ranging from a_{11} to a_{NM}.

Fig. 4 shows a more detailed block diagram of spatial multiplexing block **56.** Data **52** received by block **56** is first parsed by parser **58,** which is in direct communication with controller **66.** Based on feedback obtained from feedback extractor **80,** controller tells parser **58** into how many streams data **52** is to be divided. Parser is connected with a multiplexing block **62** and supplies the streams to its coding and modulation blocks **64.** Having separate coding and modulation blocks **64** for each stream enables the user to employ different coding rates and modulations in each stream.

Fig. 5 illustrates receive unit **90** for receiving receive signals RS from transmit unit **50** through channel **22** with N receive antennas **92.** Receive unit **90** can be any suitable receiver capable of receiving spatial multiplexed receive signals RS via the N receive antennas **92.** Exemplary receivers include maximum likelihood receivers, zero forcing equalizer receivers, successive cancellation receivers and minimum mean square error equalizer receivers. Receive unit **90** has an RF amplification and down-conversion stage **94** having individual RF amplification/down-conversion/ and analog-to-digital converter blocks **96** associated with each of the N receive antennas **72.** The N outputs of stage **94** are connected to a block **98** which performs receive processing, signal detection and decoding and demultiplexing functions. The N outputs of stage **94** are also connected to a channel estimator **100.** Channel estimator **100** obtains a measurement of channel **22;** in particular, it determines the channel

coefficients matrix H representing the action of channel **22** on transmit signals TS.

Estimator **100** is connected to block **98** to provide block **98** with the channel descriptor. The channel descriptor is typically determined by estimator **100** during training; a procedure well-known in the art. In case there is no ISI estimator **100** determines channel matrix H for the independent k streams from the k transmit antennas **72**. In case there is ISI estimator **100** determines channel matrix filter H with the aid of training sequences which are as long or longer than the delay spread. In multi-carrier operation each sub-carrier tone has a different channel so training is required for all sub-carrier tones. During training estimator **100** determines channel matrices H_1, H_2, ... H_T for all sub-carrier tones. A deinterleaver and decoder (not shown) can be placed in the data stream if a corresponding interleaver and coder was employed in transmitter **50**.

Channel estimator **100** is also connected to a mode selection block **102**. Mode selection block **102** is connected to a database **104**. Database **104** conveniently contains a look-up table similar to table 1 with quality parameters QP's and threshold values QP_{th} (i.e., the lowest acceptable values) of these QP's for each of the modes. In other words, for any particular QP each mode has an associated threshold QP_{th}, which is conveniently stored in database **104**. For example, when SINR is used as QP, then for a given (or required) performance criteria, e.g., a required BER, each

mode has a threshold $SINR_{th}$ which depends on its modulation rate and coding rate. Mode selection block **102** can thus access in database **104** the appropriate threshold QP_{th} values for the selected modes.

Alternatively, mode selection block **102** can receive quality parameter QP and threshold value QP_{th} for each mode from an outside source. In yet another embodiment, block **102** can be pre-programmed to use a particular quality parameter QP or make its own selection of quality parameter QP. Also, threshold value QP_{th} can be provided, or pre-set by block **102** or adjusted during operation by either block **102** or some other circuit, as necessary. In the present embodiment, signal-to-interference and noise ratio (SINR) is used as quality parameter QP.

Conveniently, database **104** contains the same entries as database **78** indexed by the same mode numbers. This arrangement makes it particularly easy for selection block **102** to communicate its mode selection for each transmit antenna T_i to transmit unit **50** by sending the mode number. For example, selection block **102** provides transmit antenna number and mode to be used by that transmit antenna pairwise for feedback to transmit unit **50.** In fact, antenna number and mode can be arranged in a table for feedback. When a transmit antenna number and corresponding mode are not indicated by block **102,** then that transmit antenna T_i is deactivated by controller **66.** For active transmit antennas T_i controller **66** retrieves the corresponding coding rate and modulation from database **78.** Alternatively, selection block **102** can indicate directly

which transmit antenna or antennas T_i are to be deactivated and indicate the modes to be used by active transmit antennas T_i. In some cases, the same mode can be used by all active transmit antennas T_i, e.g., at system start-up. At this time selection block **102** only sends active transmit antennas and mode for feedback to transmit unit **50.**

Mode selection block **102** is connected to a feedback block **106** for feeding back the antenna numbers and corresponding modes to receive unit **50.** Furthermore, feedback block **106** can also send channel parameters, e.g., in the form of mode number to transmit unit **50.** Receiver unit's **90** transmitter **110** is connected to feedback block **106** for transmitting this information back to transmit unit **50.**

In this embodiment, receive unit **90** is a minimum mean square error equalizer (MMSE) receiver requiring a receive processing matrix **108** based on channel matrix H to recover data **52.** Hence, mode selection block **102** has the appropriate logic to compute matrix **108** as discussed below. Block **102** communicates matrix **108** to block **98** via a link.

The operation of channel **22** on a transmit vector s of M transmit signals TS corresponding to the M transmit antennas **72** is described by the system equation:

$$x = RHs + Rv, \qquad (1)$$

where v is an Nx1 noise vector, H is the NxM channel matrix, R is the linear MMSE equalizer receiver and x is

the Nx1 receive vector estimated by receive unit **90**. It is assumed that:

$$E(ss^*) = P_o;\ E(vv^*) = R_{vv};\ E(sv^*) = 0, \qquad (2)$$

where the superscript * denotes the conjugate transpose and E is the expectation value over the distributions of v and s. The error vector e can be defined as:

$$e \equiv s - RHs - Rv. \qquad (3)$$

The linear MMSE equalizer is found by minimizing the cost function:

$$C(R) = Trace\ E(ee^*). \qquad (4)$$

Using the assumptions in (2), the cost function in (4) can be simplified to:

$$C(R) = Trace\left[P_o(I - RH)(I - RH)^* + RR_{vv}R^*\right], \qquad (5)$$

where I is the identity matrix. To obtain the optimum MMSE receiver, R_{opt}, the first derivative of the simplified cost function is set to zero, $\frac{\partial C(R)}{\partial R} = 0$, and solved for R_{opt} yielding:

$$R_{opt} = P_o H^*\left(P_o HH^* + R_{vv}\right)^{-1}. \qquad (6)$$

The receiver R, here optimized receiver R_{opt}, determines the value of quality parameter QP for transmit signals TS transmitted via each one of transmit antennas **72**. In this embodiment, signal-to-interference and noise ratio (SINR) is chosen as the quality parameter QP. The relation between $SINR_i$ for i-th of transmit antennas **72** and R_{opt} is:

$$SINR_i = \frac{P_o}{\left| C\left(R_{opt}\right) \right|_{ii}} \,. \qquad (7)$$

Of course, a person of average skill in the art will be able to construct analogous relationships between other quality parameters QP of transmit signals TS transmitted from each of transmit antennas **72** and the receiver.

In accordance with the method of the invention, a threshold value QP_{th} is assigned to quality parameter QP. The assignment of the threshold is based on the desired quality of receive signals RS. Conveniently, QP_{th} is the minimum threshold at which receive signals RS can be re-converted into data **52** by block **98** at an acceptable error rate given the mode (coding rate and modulation of data **52** employed by spatial multiplexing block **56** of transmit unit **50**). For example, the assignment can be based on a desired bit-error-rate (BER) of received data **52**. Alternatively, other error rates such as packet error rates or symbol error rates of data **52** can be used to characterize the quality of receive signals RS. A person of average skill in the art is familiar with these characterizations and criteria for their selection.

In this embodiment, data **52** is coded and modulated in accordance with a square QAM constellation (e.g. with four points, Z=4) and the quality parameter QP is SINR. The BER required at receive unit **90** given this mode assigns the minimum threshold $SINR_{th}$. Specifically, BER_i at receive unit **90** for data **52** transmitted in transmit signals TS from i-th antenna of transmit antennas **72** is related to $SINR_i$ for the i-th antenna as follows:

$$BER_i = \alpha_Z \times erfc\left(\sqrt{\beta_Z SINR_i}\right), \qquad (8)$$

where erfc is the complementary error function, $\alpha_Z = \frac{2}{\log_2 Z}\left(1 - \frac{1}{\sqrt{Z}}\right)$, and $\beta_Z = \frac{3}{2(Z-1)}$. Selecting a minimum acceptable BER_i at receive unit **90** given the mode thus yields an $SINR_i$ value to be used as threshold value $SINR_{th}$. It should be noted that this holds for uncoded schemes, i.e., when no additional coding such as error coding is imposed on data **52**. In the event such coding is used there is generally a coding gain which will vary $SINR_{th}$, as will be appreciated by a person skilled in the art.

$SINR_{th}$ is used by mode selection block **102** to identify sub-descriptors of channel **22** which do not meet threshold $SINR_{th}$. In this embodiment the sub-descriptors of channel **22** are sub-matrices h_i of channel matrix H, as shown in Fig. 6. One sub-matrix h_i, i=1...M, is associated with each transmit antenna T_i, i=1...M, of transmit antennas **72**. The schematic of Fig. 6 illustrates a case where for all sub-matrices h_i quality parameter QP exceeds threshold QP_{th} with

the exception of sub-matrix h_j. In the present embodiment QP is SINR and thus $SINR_j<SINR_{th}$.

According to the method of the invention, transmit antenna T_j corresponding to sub-matrix h_j is deactivated. In fact, transmit unit **50** is shown in Fig. 3 with antenna T_j deactivated, i.e., no transmit signals TS are transmitted from antenna T_j. This deactivation of one or more transmitting antennas T_i will generally improve and not worsen the quality parameter QP, in this case SINR, for the remaining transmitting antennas.

Figs. 7 illustrates how mode selection block **102** implements the deactivation decision. Channel matrix H, QP and QP_{th} are received by a comparison block **110** where QP values for each sub-matrix h_i are compared with QP_{th} given the selected mode (see Fig. 6). Comparison block **110** identifies which of sub-matrices h_i has a QP_i less than QP_{th} and removes this sub-matrix, in the present embodiment sub-matrix h_j, since $QP_j<QP_{th}$, from channel matrix H. Removal of sub-matrix h_j from channel matrix H produces a subset channel matrix H'. Comparison block **110** recomputes QP_i for each sub-matrix h_i corresponding to transmit antenna T_i in subset matrix H'. Conveniently, recomputed QP_i of sub-matrices h_i of subset channel matrix H' are compared with adjusted QP_{th}. For example, recomputed QP_i are compared with QP_{th} required for particular modes to determine in which of those modes the corresponding antennas T_i should transmit. Advantageously, the mode whose required QP_{th} is closest in value to the recalculated QP_i is selected for data **52** transmitted from corresponding transmit antenna T_i.

Block **110** also passes subset channel matrix H' to a computing block **112.** Computing block **112** calculates the processing matrix **108** or optimal receiver R_{opt} and sends R_{opt} to block **98** for receive processing.

When all transmit antennas T_i meet threshold QP_{th} of a different mode then employed at the time, e.g., a mode with a higher throughput, then this higher throughput mode is selected by comparison block **110.** When a sub-matrix h_j indicates that the corresponding transmit antenna T_j no longer meets QP_{th}, then that antenna is deactivated and the mode is fed back to transmit unit **50.** In this manner, transmission of data **52** can be optimized for highest throughput at the set QP_{th}.

In fact, flow chart of Fig. 8 shows an embodiment of the method for achieving highest throughput a mode number # with corresponding threshold $QP_{th-\#}$. Initially, i is set equal to the number of transmit antennas T_i, i=M, such that k=M. Then, optimal receiver R_{opt} is computed as well as the values QP_i for all sub-matrices h_i. For convenience, antennas T_i and their corresponding sub-matrices h_i are arranged in descending order of QP_i. A mapping of this re-arranged or ordered set to the original order of sub-matrices h_i and corresponding transmit antennas T_i is maintained for administrative purposes.

In the next step, each QP_i is compared with a lowest threshold $QP_{th-\#}$. For example, lowest threshold $QP_{th-\#}$ can be equal to the threshold for the lowest throughput acceptable

mode #. When a QP_i does not meet this lowest threshold $QP_{th-\#}$, the corresponding sub-matrix h_i is removed from channel matrix H to produce subset channel matrix H'. With the same action the corresponding antenna T_i is designated for deactivation. After removal of sub-matrix h_i the values of QP_i are re-computed and the comparison repeated, until subset channel matrix H' contains only sub-matrices h_i which have QP_i higher than $QP_{th-\#}$. It should be noted that more than one sub-matrix h_i can be removed at a time.

Once the final subset channel matrix H' is obtained it is sent to the second branch in the flow chart of Fig. 8 to determine the best modes to use for transmission from the remaining antennas. The number of data streams k is set to the number of remaining sub-matrices h_i, i=k. Then, in a recursive loop process, the best mode number for each antenna is determined by direct comparison of QP_k with $QP_{th-\#}$ required for that mode #. The mode # for which the comparison yields the closest match is selected for transmission from corresponding antenna #. Conveniently, the antenna # and mode # are ordered pairwise in a table for feedback to the transmit unit. Before feedback, the table is arranged to agree with the updated mapping of antenna # which was performed to arrange sub-matrices h_i in descending order of QP_i.

It should be noted, that comparison block **110** can re-set or adjust lowest $QP_{th-\#}$. For example, when data **52** is not very sensitive (e.g., voice) lowest $QP_{th-\#}$ can be lowered and when data **52** is sensitive lowest threshold $QP_{th-\#}$ can be raised. In fact, the setting of lowest $QP_{th-\#}$ depends on the type of

data **52** and other parameters well-known in the art of data processing.

Comparison block **110** can repeat the steps of identifying individual or even groups or sets of sub-matrices h_i falling below lowest $QP_{th-\#}$ and deactivate the corresponding antenna or antennas among transmit antennas **72**. Of course, when channel **22** is very high quality, no transmit antennas **72** may need to be deactivated.

In general, the time period within which the above computations for antenna deactivation should be repeated should be shorter than the coherence time of channel **22**.

In another embodiment, transmit unit **50** receiving feedback of channel information, whether using TDD or simple feedback, could make the selection of antenna or antennas to deactivate on its own. This alternative approach would be convenient when receive unit **90** does not have sufficient resources or power to make the comparisons between the values of QP_i and QP_{th}. Of course, transmit unit **50** would then contain all the corresponding computation and decision-making blocks, specifically mode selection block **102**, contained in receive unit **90** as described above.

In an alternative embodiment shown in Fig. 9, the method of invention is employed in a communications system **200** using spatial-multiplexing. System **200** has several base transceiver stations (BTS), of which two **204, 206** are shown. BTS **204, 206** are equipped with transmit antenna arrays **208, 210** respectively for sending transmit signals

to receive unit **212**. It should be noted, however, that a combination of BTS with single transmit antennas can be used as long as spatial multiplexing is employed.

Receive unit **212** sets a threshold QP_{th} and identifies among the sub-descriptors of a descriptor of channel **22**, e.g., h_i matrices of channel matrix H, one or more sub-descriptors which do not meet threshold QP_{th} in the selected mode. As described above, receive unit **212** can adjust threshold QP_{th}, in particular, it can adjust threshold QP_{th} based on the desired mode. Receive unit **212** then determines a final selection which transmit antennas of transmit antenna arrays **208, 210** should be deactivated. Additionally, receive unit **212** determines which modes should be used by the remaining active transmit antennas of arrays **208, 210**. It should be noted that under certain circumstances receive unit **212** may determine that one of base stations **204, 206** should not be transmitting any transmit signals to receive unit **212** at all.

The selection of antennas and modes is fed back from receive unit **212** to BTS **204, 206**. In particular, a control logic, in this embodiment a controller **202** receives the feedback from receive unit **212**. Controller **202** can be a central control unit supervising the operation of BTS **204, 206** and any other BTS of communication system **200**. Alternatively, control logic can consist of separate control units as indicated in dashed lines.

In another embodiment of the invention, the descriptor of channel **22** is a channel matrix filter **H** and sub-descriptors are sub-matrix filters $\mathbf{h}_i$. Conveniently, channel matrix filter **H** is used as descriptor when inter-symbol interference (ISI) is present due to broadly varying times of arrival or under other adverse conditions associated with delay spread. A person of average skill in the art is familiar with channel matrix filters **H**, their construction and their use in reconstructing transmitted data under such channel conditions.

In particular, for a channel with delay spread the system equation is:

$$X = \mathcal{H}S + N \,. \tag{9}$$

This equation can be rewritten in matrix form as:

$$\begin{pmatrix} \underline{x}_k \\ \underline{x}_{k-1} \\ \cdot \\ \cdot \\ \cdot \\ \underline{x}_{k-q+1} \end{pmatrix} = \begin{array}{rc} \begin{array}{c} 1 \\ 2 \\ \cdot \\ \cdot \\ \cdot \\ q \end{array} & \begin{pmatrix} H_o & H_1 \ldots H_L & 0 \ldots & 0 \\ 0 & H_o & H_1 \ldots H_L & 0 \\ \cdot & & & \\ \cdot & \cdot & \cdot & \cdot \\ \cdot & & & \\ 0 & 0 & .. & H_o \ldots H_L \end{pmatrix} \\ & \begin{array}{cccc} 1 & 2 & \ldots. & p \end{array} \end{array} \begin{pmatrix} \underline{s}_k \\ \underline{s}_{k-1} \\ \cdot \\ \cdot \\ \cdot \\ \underline{s}_{k-p+1} \end{pmatrix} + \begin{pmatrix} \underline{n}_k \\ \underline{n}_{k-1} \\ \cdot \\ \cdot \\ \cdot \\ \underline{n}_{k-p+1} \end{pmatrix},$$

where $\underline{s}_k$ is the signal vector at time $\underline{k}$, $\underline{n}_k$ is the noise vector at time $\underline{k}$, $\underline{x}_k$ is the received signal vector at time $\underline{k}$, q is the number of received signal snap shots used for processing (also referred to as the number of equalizer taps), L is the number of delay elements in the channel and

$p \geq q+L$. As is clear from the explicit system equation, channel matrix filter $\mathcal{H}$ actually consists of a number of channel matrices H, each of which is an NxM matrix, where N is the number of receive antennas at the receive unit and M is the number of transmit antennas. Therefore, channel matrix filter H is actually qNxpM. In fact, M is the total number of antennas from which receive unit is meant to receive spatially multiplexed signals. As discussed above, these transmit antennas can belong to separate BTS, include single transmit antennas or transmit antennas which are part of any suitable transmit antenna array.

The receiver equalizer R(z) for processing the q time snap shots of the received vector $\underline{x}_k$ to obtain a good MMSE estimator of the transmitted signal vector $\underline{s}_k$ can be represented as follows:

$$R(z) = \sum_{i=0}^{q-1} R_i z^{-i} \quad , \tag{10}$$

where z is the delay element, R_i is the equalizer tap at the i-th instant. Using system equation (9) and equation (10) the recovered signal vector, $\hat{\underline{s}}_k$, where the hat indicates recovered signal, can be written as:

$$\hat{\underline{s}}_k = [R_o, \mathrm{R}_1 \ldots R_{q-1}]\mathcal{H}S + [R_o, \mathrm{R}_1 \ldots R_{q-1}]N \, . \tag{11}$$

By defining the following correlation matrices:

$$E[NN^*] = \sigma^2 I \tag{12}$$

$$E[SS^*] = I, \tag{13}$$

where σ^2 is the noise variance and I is the identity matrix, the MMSE estimator R for estimating $\hat{\underline{s}}_k$ from S can be written as:

$$R = [H_o^* \, 0 \, 0 \, \ldots \,](\mathcal{H}\mathcal{H}^* + \sigma^2)^{-1}. \tag{14}$$

In this embodiment the quality parameter QP is chosen to be signal-to-noise ratio (SNR). Now, for the i-th transmit antenna the SNR is given by:

$$SNR_i = \frac{1}{\sigma^2}[R\mathcal{H}\mathcal{H}^* R^*]_{ii}, \tag{15}$$

where i=1, 2, ... M. At this point the SNR values for each transmit antenna can be compared with a threshold SNR_{th}. The remaining steps leading to the selection of which transmit antenna should be deactivated are analogous to those described above.

The method of invention can also be used in multi-carrier systems using spatial multiplexing. In these types of communication systems several sub-carrier frequencies or sub-carrier tones are transmitted from each transmit antenna. Each of these sub-carrier tones experiences a different channel in propagating from the transmit unit to the receive unit. Hence, each transmit antenna has associated with it a group of sub-descriptors; one for each sub-carrier tone.

In this case each transmit antenna and its associated sub-descriptor will yield a different quality parameter QP at the different sub-carrier frequencies. In other words, data transmitted from a transmit antenna T_j on a first sub-carrier tone ST_1 will exhibit a quality parameter QP_{j1} different from a quality parameter QP_{j2} for data transmitted from the same transmit antenna T_j on a second sub-carrier tone ST_2. Hence, the selection of which transmit antenna to deactivate in this case is made based on the average quality parameter $\underline{QP}_j$ for the group of sub-descriptors associated with that antenna. With this change, the method of invention can be employed in multi-carrier systems as described for non multi-carrier communication systems.

The method of the invention can be used in soft hand-off between base stations in cellular systems. Alternatively, partial hand-offs or gradual hand-offs between base stations can be implemented, depending on antenna selection. In fixed wireless systems the user device can operate by receiving signals from a selection of antennas belonging to different base stations at all times. Also, in the case of fixed wireless devices there may be cases where the channel does not change appreciably over time. In this case, once the antenna selection is made, no feedback may be required.

It will be clear to one skilled in the art that the above embodiments may be altered in many ways without departing from the scope of the invention. Accordingly, the scope of

the invention should be determined by the following claims and their legal equivalents.

CLAIMS

What is claimed is:

1. A method for selecting antennas for spatial multiplexed communication in a channel for transmitting data between a transmit unit having transmit antennas and a receive unit having receive antennas, said method comprising:
   a) determining a channel descriptor having sub-descriptors corresponding to said transmit antennas;
   b) choosing a quality parameter;
   c) assigning a threshold to said quality parameter;
   d) identifying among said sub-descriptors a sub-descriptor not meeting said threshold; and
   e) deactivating among said transmit antennas an antenna associated with said sub-descriptor.

2. The method of claim 1, wherein said quality parameter is selected from the group consisting of signal-to-interference and noise ratio, signal-to-noise ratio, power level.

3. The method of claim 1, wherein said channel descriptor is a channel matrix filter H, said sub-descriptors are sub-matrix filters h_i of said channel matrix filter H, and said sub-descriptor is a sub-matrix filter h_j.

4. The method of claim 3, further comprising removing said sub-matrix filter h_j from said channel matrix filter H to obtain a subset channel matrix filter H'.

5. The method of claim 4, further comprising repeating said identifying and said deactivating.

6. The method of claim 1, wherein said spatial multiplexed communication employs a number of sub-carrier tones associating groups of said sub-descriptors with said transmit antennas.

7. The method of claim 6, wherein said identifying comprises computing an average quality parameter for said groups of sub-descriptors and identifying among said groups of sub-descriptors a group of sub-descriptors not meeting said threshold, and said deactivating comprises deactivating among said transmit antennas an antenna associated with said group of sub-descriptors.

8. The method of claim 1, wherein said channel descriptor is a channel matrix H, said sub-descriptors are sub-matrices h_i of said channel matrix H, and said sub-descriptor is a sub-matrix h_j.

9. The method of claim 8, further comprising removing said sub-matrix h_j from said channel matrix H to obtain a subset channel matrix H'.

10. The method of claim 9, further comprising repeating said identifying and said deactivating.

11. The method of claim 1, further comprising adjusting said threshold.

12. The method of claim 1, wherein said data is coded and modulated in accordance with a selected mode.

13. The method of claim 12, wherein said mode comprises a predetermined coding rate and modulation.

14. The method of claim 12, wherein said threshold is assigned based on said selected mode.

15. The method of claim 12, wherein said selected mode is based on said quality parameter.

16. The method of claim 12, wherein said selected mode is fed back to said transmit unit.

17. The method of claim 1, wherein said receive unit is selected from the group consisting of maximum likelihood receivers, zero forcing equalizer receivers, successive cancellation receivers and minimum mean square error equalizer receivers.

18. A method for selecting antennas for spatial multiplexed communication in a channel for transmitting data between a transmit unit having transmit antennas and a receive unit having receive antennas, said method comprising:

a) determining a channel descriptor having sub-descriptors corresponding to said transmit antennas;
b) choosing a quality parameter;
c) assigning a threshold to said quality parameter;
d) identifying among said sub-descriptors a set of sub-descriptors not meeting said threshold; and
e) deactivating among said transmit antennas, antennas associated with said set of sub-descriptors.

19. The method of claim 18, wherein said quality parameter is selected from the group consisting of signal-to-interference and noise ratio, signal-to-noise ratio, power level.

20. The method of claim 18, wherein said channel descriptor is a channel matrix filter H, said

sub-descriptors are sub-matrix filters h_i of said channel matrix filter H, and said set of sub-descriptors is a set of said sub-matrix filters h_i.

21. The method of claim 20, further comprising removing said set of sub-matrix filters h_i from said channel matrix filter H to obtain a subset channel matrix filter H'.

22. The method of claim 21, further comprising repeating said identifying and said deactivating.

23. The method of claim 18, wherein said spatial multiplexed communication employs a number of sub-carrier tones associating groups of said sub-descriptors with said transmit antennas.

24. The method of claim 23, wherein said identifying comprises computing an average quality parameter for said groups of sub-descriptors and identifying among said groups of sub-descriptors a set of groups of sub-descriptors not meeting said threshold, and said deactivating comprises deactivating among said transmit antennas a set of antennas associated with said set of groups of sub-descriptors.

25. The method of claim 18, wherein said channel descriptor is a channel matrix H, said sub-descriptors are sub-matrices h_i of said channel matrix H, and said set of sub-descriptors is a set of said sub-matrices h_i.

26. The method of claim 25, further comprising removing said set of sub-matrices h_i from said channel matrix H to obtain a subset channel matrix H'.

27. The method of claim 26, further comprising repeating said identifying and said deactivating.

28. The method of claim 18, further comprising adjusting said threshold.

29. The method of claim 18, wherein said data is coded and modulated in accordance with a selected mode.

30. The method of claim 29, wherein said mode comprises a predetermined coding rate and modulation.

31. The method of claim 29, wherein said threshold is assigned based on said selected mode.

32. The method of claim 29, wherein said selected mode is based on said quality parameter.

33. The method of claim 29, wherein said selected mode is fed back to said transmit unit.

34. The method of claim 18, wherein said receive unit is selected from the group consisting of maximum likelihood receivers, zero forcing equalizer receivers, successive cancellation receivers and minimum mean square error equalizer receivers.

35. A method for selecting antennas for spatial multiplexed communication in a channel for transmitting data between transmit antennas and a receive unit having an array of receive antennas, said method comprising:

a) determining a channel descriptor having sub-descriptors corresponding to said transmit antennas;
b) choosing a quality parameter;
c) assigning a threshold to said quality parameter;
d) identifying among said sub-descriptors a sub-descriptor not meeting said threshold; and
e) deactivating among said transmit antennas an antenna associated with said sub-descriptor.

36. The method of claim 35, wherein said quality parameter is selected from the group consisting

of signal-to-interference and noise ratio, signal-to-noise ratio, power level.

37. The method of claim 35, wherein said channel descriptor is a channel matrix filter $\mathbf{H}$, said sub-descriptors are sub-matrix filters h_i of said channel matrix filter $\mathbf{H}$, and said sub-descriptor is a sub-matrix filter h_j.

38. The method of claim 37, further comprising removing said sub-matrix filter h_j from said channel matrix filter $\mathbf{H}$ to obtain a subset channel matrix filter $\mathbf{H}'$.

39. The method of claim 38, further comprising repeating said identifying and said deactivating.

40. The method of claim 35, wherein said spatial multiplexed communication employs a number of sub-carrier tones associating groups of said sub-descriptors with said transmit antennas.

41. The method of claim 40, wherein said identifying comprises computing an average quality parameter for said groups of sub-descriptors and identifying among said groups of sub-descriptors a group of sub-descriptors not meeting said threshold, and

said deactivating comprises deactivating among said transmit antennas an antenna associated with said group of sub-descriptors.

42. The method of claim 35, wherein said channel descriptor is a channel matrix H, said sub-descriptors are sub-matrices h_i of said channel matrix H, and said sub-descriptor is a sub-matrix h_j.

43. The method of claim 42, further comprising removing said sub-matrix h_j from said channel matrix H to obtain a subset channel matrix H'.

44. The method of claim 43, further comprising repeating said identifying and said deactivating.

45. The method of claim 35, further comprising adjusting said threshold.

46. The method of claim 35, wherein said selected mode is based on said quality parameter.

47. The method of claim 35, wherein said receive unit is selected from the group consisting of maximum likelihood receivers, zero forcing equalizer receivers, successive cancellation receivers and minimum mean square error equalizer receivers.

48. A system for spatial multiplexed communication in a channel for transmitting data between a transmit unit having transmit antennas and a receive unit having receive antennas, said receive unit comprising:

a) a channel estimation block for determining a channel descriptor having sub-descriptors corresponding to said transmit antennas;

b) a mode selection block for receiving a quality parameter, assigning a threshold to said quality parameter, and identifying among said sub-descriptors at least one sub-descriptor not meeting said threshold;

c) a feedback unit for sending feedback related to said at least one sub-descriptor to said transmit unit;

and said transmit unit comprising a control logic for receiving said feedback and deactivating among said transmit antennas, antennas associated with said at least one sub-descriptor.

49. The system of claim 48, wherein said transmit unit further comprises a spatial multiplexing block connected to said controller for coding and multiplexing said data.

50. The system of claim 48, wherein said transmit unit further comprises a switching unit for deactivating said antennas.

51. The system of claim 48, wherein said receive unit is selected from the group consisting of maximum likelihood receivers, zero forcing equalizer receivers, successive cancellation receivers and minimum mean square error equalizer receivers.

52. A system for spatial multiplexed communication in a channel for transmitting data between a number of transceivers having transmit antennas and a receive unit having receive antennas, said receive unit comprising:

a) a channel estimation block for determining a channel descriptor having sub-descriptors corresponding to said transmit antennas;

b) a mode selection block for receiving a quality parameter, assigning a threshold to said quality parameter, and identifying among said sub-descriptors at least one sub-descriptor not meeting said threshold;

c) a feedback unit for sending feedback related to said at least one sub-descriptor to said transmit unit;

and said number of transceivers comprising a control logic for receiving said feedback and deactivating among said transmit antennas, antennas associated with said at least one sub-descriptor.

53. The system of claim 52, wherein said transceivers further comprise spatial multiplexing blocks for coding and multiplexing said data.

54. The system of claim 52, wherein said transceivers further comprise switching units, for deactivating said antennas.

55. The system of claim 52, wherein said receive unit is selected from the group consisting of maximum likelihood receivers, zero forcing equalizer receivers, successive cancellation receivers and minimum mean square error equalizer receivers.

56. The system of claim 52, wherein said control logic comprises a number of control units.

57. The system of claim 56, wherein each of said transceivers has one of said control units.

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US01/11309

A. CLASSIFICATION OF SUBJECT MATTER
IPC(7) : H04B 1/38
US CL : 455/63, 562; 370/329
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
U.S. : 455/63, 69, 272, 562; 370/329, 334

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)
Please See Continuation Sheet

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category *	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US 4,797,947 A (LABEDZ) 10 January 1989, abstract.	1-57
A	US 5,471,647 A (GERLACH et al) 28 November 1995, col. 3, line 6 - col. 6, line 35.	1-57
A	US 5,592,490 A (BARRATT et al) 07 January 1997, abstract.	1-57

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

* Special categories of cited documents:

"A" document defining the general state of the art which is not considered to be of particular relevance

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"P" document published prior to the international filing date but later than the priority date claimed

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"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

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"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
12 June 2001 (12.06.2001)	23 JUL 2001

Name and mailing address of the ISA/US	Authorized officer
Commissioner of Patents and Trademarks Box PCT Washington, D.C. 20231	LEE NGUYEN Karen L. Ward
Facsimile No. (703)305-3230	Telephone No. 703-305-4700

Form PCT/ISA/210 (second sheet) (July 1998)

Continuation of B. FIELDS SEARCHED Item 3: EAST
search term: spatial division multiple access, multiplex, antenna, deactivate, feedback

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
28 March 2002 (28.03.2002)

PCT

(10) International Publication Number
WO 02/25853 A2

(51) International Patent Classification[7]: H04L

(21) International Application Number: PCT/US01/29368

(22) International Filing Date:
19 September 2001 (19.09.2001)

(25) Filing Language: English

(26) Publication Language: English

(30) Priority Data:
09/665,149 19 September 2000 (19.09.2000) US

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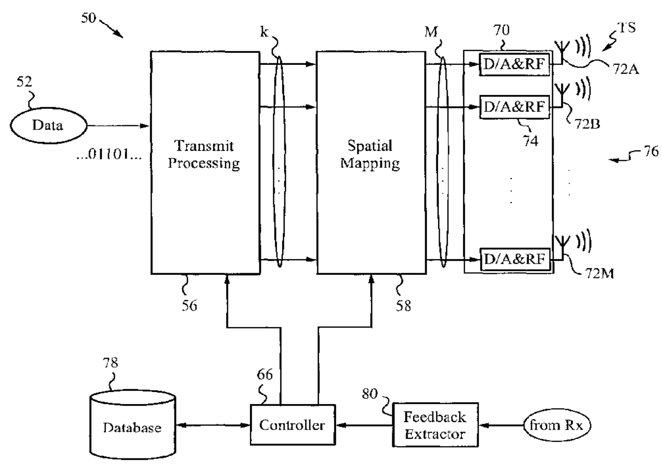
(81) Designated States *(national)*: AE, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, CA, CH, CN, CU, CZ, DE, DK, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MD, MG, MK, MN, MW, MX, NO, NZ, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TR, TT, UA, UG, UZ, VN, YU, ZA, ZW.

(84) Designated States *(regional)*: ARIPO patent (GH, GM, KE, LS, MW, MZ, SD, SL, SZ, TZ, UG, ZW), Eurasian patent (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European patent (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE, TR), OAPI patent (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

Published:
— *without international search report and to be republished upon receipt of that report*

[Continued on next page]

(54) **Title:** MODE SELECTION FOR DATA TRANSMISSION IN WIRELESS COMMUNICATION CHANNELS BASED ON STATISTICAL PARAMETERS



(57) **Abstract:** A method and communication system for selecting a mode for encoding data for transmission in a wireless communication channel between a transmit unit and a receive unit. The data is initially transmitted in an initial mode and the selection of the subsequent mode is based on a selection of first-order and second-order statistical parameters of short-term and long-term quality parameters. Suitable short-term quality parameters include signal-to-interference and noise ratio (SINR), signal-to-noise ratio (SNR), power level and suitable long-term quality parameters include error rates such as bit error rate (BER) and packet error rate (PER). The method of the invention can be employed in Multiple Input Multiple Output (MIMO), Multiple Input Single Output (MISO), Single Input Single Output (SISO) and Single Input Multiple Output (SIMO) communication systems to make subsequent mode selection faster and more efficient. Furthermore the method can be used in communication systems employing various transmission protocols including OFDMA, FDMA, CDMA, TDMA.

WO 02/25853 A2

Mode Selection for Data Transmission in Wireless Communication Channels Based on Statistical Parameters

FIELD OF THE INVENTION

The present invention relates generally to wireless communication systems and methods, and more particularly to mode selection for encoding data for transmission in a wireless communication channel based on statistical parameters.

BACKGROUND OF THE INVENTION

Wireless communication systems serving stationary and mobile wireless subscribers are rapidly gaining popularity. Numerous system layouts and communications protocols have been developed to provide coverage in such wireless communication systems.

Wireless communications channels between transmit and receive devices are inherently variable and their quality fluctuates. Specifically, the quality parameters of such communications channels vary in time. Under good conditions wireless channels exhibit good communication parameters, e.g., large data capacity, high signal quality, high spectral efficiency and throughput. At these times significant amounts of data can be transmitted via the channel reliably. However, as the channel changes in time, the communication parameters also change. Under altered conditions former data rates, coding techniques and data formats may no longer be feasible. For example, when channel performance is degraded the transmitted data may experience excessive corruption yielding unacceptable communication parameters. For instance, transmitted data can exhibit excessive bit-error rates or packet error rates. The degradation of the channel can be due to a multitude of factors such as general noise in the channel, multi-path fading, loss

of line-of-sight path, excessive Co-Channel Interference (CCI) and other factors.

In mobile systems, signal degradation and corruption is chiefly due to interference from other cellular users within or near a given cell and multipath fading, in which the received amplitude and phase of a signal varies over time. The fading rate can reach as much as 200 Hz for a mobile user traveling at 60 mph at PCS frequencies of about 1.9 GHz. In such environments, the problem is to cleanly extract the signal of the user being tracked from the collection of received noise, CCI, and desired signal portions.

In Fixed Wireless Access (FWA) systems, e.g., where the receiver remains stationary, signal fading rate is less than in mobile systems. In this case, the channel coherence time or the time during which the channel estimate remains stable is longer since the receiver does not move.

Prior art wireless systems have employed adaptive modulation of the transmitted signals with the use of feedback from the receiver as well as adaptive coding and receiver feedback to adapt data transmission to changing channel conditions. Such adaptive modulation is applied to Single Input Single Output (SISO) and Multiple Input Multiple Output (MIMO) systems, e.g., systems with antenna arrays at the transmit and receive ends.

In both SISO and MIMO systems, however, the fundamental problem of efficient choice of the mode to be applied to the transmitted data remains. For general prior art on the subject the reader is referred to A.J. Goldsmith et al., "Variable-rate variable power MQAM for fading channels", IEEE Transactions of Communications, Vol. 45, No. 10, Oct. 1997, pp. 1218-1230; P.

Schramm et al., "Radio Interface of EDGE, a proposal for enhanced data rates in existing digital cellular systems", Proceedings IEEE 48th Vehicular Technology Conference (VTC' 1998), pp. 1064-1068; and Van Noblen et al., "An adaptive link protocol with enhanced data rates for GSM evolution", IEEE Personal Communications, February 1999, pp. 54-63.

U.S. Pat. No. 6,044,485 to Dent et al. teaches a transmission method and system which adapts the coding of data based on channel characteristics. The channel characteristics are obtained either from a channel estimation circuit or from an error feedback signal. In particular, when the signal-to-noise (SNR) characteristic is chosen as an indicator of the state of the channel, then a different coding is applied to the data being transmitted for high and low SNR states of the channel. In addition, the encoding also employs information derived from the cyclic redundancy check (CRC).

The method taught by Dent varies the coding rate only and not the modulation rate. This method does not permit one to select rapidly and efficiently from a large number of encoding modes to adapt to varying channel conditions.

U.S. Pat. No. 5,559,810 to Gilbert et al. teaches a communication system using data reception history for selecting a modulation technique from among a plurality of modulation techniques to thus optimize the use of communication resources. At least one block of data is transmitted with a particular modulation technique and a data reception history is maintained to indicate transmission errors, e.g., by keeping a value of how many blocks had errors. The data reception history is updated and used to determine an estimate of transmission

signal quality for each modulation technique. This value is then used in selecting the particular modulation technique.

The system and method taught by Gilbert rely on tracking errors in the reception of entire blocks of data. In fact, signal quality statistics, e.g., signal-to-interference and noise ratio (SINR) are derived from the error numbers for entire blocks of data, which requires a significant amount of time. Thus, this method does not permit one to select rapidly and efficiently from a large number of encoding modes to adapt to varying channel conditions, especially in the event of rapid fades as encountered in mobile wireless systems.

It would be an advance to provide a mode selection technique which allows the system to rapidly and efficiently select the appropriate mode for encoding data in a quickly changing channel. It is important that such technique be efficient in all wireless systems, including Multiple Input Multiple Output (MIMO), Multiple Input Single Output (MISO), Single Input Single Output (SISO) and Single Input Multiple Output (SIMO) systems as well as systems using multiple carrier frequencies, e.g., OFDM systems.

SUMMARY

The present invention provides a method for selecting a mode for encoding data for transmission in a wireless communication channel between a transmit unit and a receive unit. The data is first encoded in accordance with an initial mode and transmitted from the transmit unit to the receive unit. One or more quality parameters are sampled in the data received by the receive unit. Then, a first-order statistical parameter and a second-order statistical parameter of the quality parameter are

computed and used for selecting a subsequent mode for encoding the data.

The one or more quality parameters can include a short-term quality parameter or several short-term quality parameters and be selected among parameters such as signal-to-interference and noise ratio (SINR), signal-to-noise ratio (SNR) and power level. Conveniently, a first sampling time or window is set during which the short-term quality parameter is sampled. In one embodiment, the length of the first sampling window is based on a coherence time of the wireless communication channel. In another embodiment, the duration of the first sampling window is based on a delay time required to apply the subsequent mode to encode the data at the transmit unit. In yet another embodiment, the second-order statistical parameter is a variance of the short-term quality parameter and the length of the first sampling window is selected on the order of the variance computation time.

The one or more quality parameters can also include a long-term quality parameter or several long-term quality parameters. The long-term quality parameter can be an error rate of the data, such as a bit error rate (BER) or a packet error rate (PER) at the receive unit. Again, it is convenient to set a second sampling time or window during which the long-term quality parameter is sampled. In one embodiment, the first-order statistical parameter is a mean of the long-term quality parameter and the length of the second sampling window is set on the order of the mean computation time. In another embodiment, the length of the second sampling window is set on the order of an error rate computation time.

In many instances, it is convenient when the first-order statistical parameter is a mean of the quality parameter and the second-order statistical parameter is a variance of the quality parameter. The variance can include two variance types: a temporal variance and a frequency variance. The latter is useful in systems employing multiple frequencies for transmitting the data. Specifically, it is particularly convenient to compute both temporal and frequency variances when the data is transmitted in accordance with a multi-carrier scheme.

The initial mode for encoding the data can be selected from a set of modes. The set of modes can be made up of a number of modes which are likely to work based on a preliminary analysis of the channel. The set of modes can be organized in accordance with the at least one quality parameter whose first-order and second-order statistics are used in subsequent mode selection.

Conveniently, the subsequent mode is communicated to the transmit unit and applied to the data to maximize a communication parameter in the channel. For example, the subsequent mode can maximize data capacity, signal quality, spectral efficiency or throughput of the channel or any other communication parameter or parameters as desired.

The method of the invention can be used in Multiple Input Multiple Output (MIMO), Multiple Input Single Output (MISO), Single Input Single Output (SISO) and Single Input Multiple Output (SIMO) communication systems, e.g., receive and transmit units equipped with multiple antennas. Furthermore the method can be used in communication systems employing various transmission protocols including OFDMA, FDMA, CDMA, TDMA.

The method of invention can also be used for selecting the mode from a set of modes and adjusting the selection. For this purpose data encoded in an initial mode selected from the set of modes is received by the receive unit. The short-term quality parameter is then sampled to determine a statistical parameter of the short-term quality parameter. Of course, the statistical parameter can include any combination of first-order and second-order statistical parameters. The subsequent mode is selected based on the short-term statistical parameter. In addition, the long-term quality parameter of the data received by the receive unit is also sampled. The subsequent mode selected based on the short-term statistical parameter is then adjusted based on the long-term quality parameter.

The set of modes can be arranged in any suitable manner, e.g., it can be arranged in a lookup table and ordered by the short-term quality parameter and specifically the first-order and/or second-order statistics of the short-term quality parameter for easy selection. In fact, the lookup table can be modified based on the short-term quality parameter.

The invention also encompasses a system for assigning a subsequent mode for encoding data. The system has a transmit unit equipped with a transmit processing block for encoding the data in a mode. A receive unit is provided for receiving the data transmitted from the transmit unit. The receive unit has a statistics computation block for sampling at least one quality parameter of the received data and computing the first-order and second-order statistical parameters of the data. The receive unit also has a mode selection block for assigning the subsequent mode based on the first-order and second-order statistical parameters.

Conveniently, the system has at least one database containing the set of modes from which the mode, e.g., the initial mode, and the subsequent mode are selected. In one case, the receive has a first database containing the modes and the transmit unit has a second such database.

The system also has a feedback mechanism for communicating the subsequent mode from the receive unit to the transmit unit. This feedback mechanism can be a separate mechanism or comprise the time-division duplexing (TDD) mechanism.

A detailed description of the invention and the preferred and alternative embodiments is presented below in reference to the attached drawing figures.

BRIEF DESCRIPTION OF THE FIGURES

Fig. 1 is a simplified diagram illustrating a communication system in which the method of the invention is applied.

Fig. 2 is a graph illustrating the effects of channel variation in time and frequency.

Fig. 3 is a block diagram of an exemplary transmit unit in accordance with the invention.

Fig. 4 is a block diagram of an exemplary receive unit in accordance with the invention.

Fig. 5 is a schematic diagram illustrating data transmitted in a wireless communication channel.

DETAILED DESCRIPTION

The method and systems of the invention will be best understood after first considering the simplified diagram of Fig. 1

illustrating a portion of a wireless communication system **10**, e.g., a cellular wireless system in which the method of invention can be employed. For explanation purposes, downlink communication will be considered where a transmit unit **12** is a Base Transceiver Station (BTS) and a receive unit **14** is a mobile or stationary wireless user device. Of course, the method can be used in uplink communication from receive unit **14** to BTS **12**.

Exemplary user devices **14** include mobile receive units such as a portable telephone **14A**, a car phone **14B** and a stationary receive unit **14C**. Receive unit **14C** can be a wireless modem used at a residence or any other fixed wireless unit. Receive units **14A** and **14C** are equipped with multiple antennas or antenna arrays **20**. These receive units can be used in Multiple Input Multiple Output (MIMO) communications taking advantage of techniques such as spatial multiplexing or antenna diversity. Receive unit **14B** has a single antenna **19** and can be used in Single Input Single Output (SISO) communications. It will be understood by those skilled in the art that receive units **14A**, **14B**, **14C**, could be equipped in SISO, MISO (Multiple Input Single Output), SIMO (Single Input Multiple Output), or MIMO configurations. For example, in Fig. 1 receive unit **14B** is shown having a single antenna therefore it can be employed in SISO or MISO configurations. MISO configuration can be realized in the case of **14B** for example by receiving signals from the antenna array at BTS **12A** or from distinct BTS such as **12B**, or any combination thereof. With the addition of multiple receive antennas **14B**, as **14A** and **14C**, could also be used in SIMO or MIMO configurations. In any of the configurations discussed above, the communications techniques can employ single-carrier or multi-carrier communications techniques.

A first exemplary transmit unit **12** is a BTS **12A** equipped with an antenna array **16** consisting of a number of transmit antennas **18A, 18B, ..., 18M** for MIMO communications. Another exemplary transmit unit **12** is a BTS **12B** equipped with a single omnidirectional antenna **13.** BTSs **12A, 12B** send data in the form of transmit signals TS to receive units **14A, 14B, 14C** via wireless communications channels **22.** For simplicity, only channel **22A** between BTS **12A** and receive unit **14A** and channel **22B** between BTS **12B** and receive unit **14C** are indicated.

The time variation of channels **22A, 22B** causes transmitted signal TS to experience fluctuating levels of attenuation, interference, multi-path fading and other deleterious effects. Therefore, communication parameters of channel **22A** such as data capacity, signal quality, spectral efficiency and throughput undergo temporal changes. The cumulative effects of these variations of channel **22A** between BTS **12A** and receive unit **14A** are shown for illustrative purposes in Fig. 2. In particular, this graph shows the variation of a particular quality parameter, in this case signal strength of receive signal RS at receive unit **14A** in dB as a function of transmission time t and frequency f of transmit signal TS sent from transmit unit **12A.** Similar graphs can be obtained for other quality parameters, such as signal-to-interference and noise ratio (SINR), signal-to-noise ratio (SNR) as well as any other quality parameters known in the art. Of the various quality parameters signal strength (power level), SINR and SNR are generally convenient to use because they can be easily and rapidly derived from receive signals RS as is known in the art.

In accordance with the invention, a mode for encoding data at transmit units **14** is selected based on a first order statistical parameter and a second order statistical parameter

of the quality parameter. The details of the method will now be explained by referring to the operation of a transmit unit **50**, as illustrated in Fig. 3 and a receive unit **90** as illustrated in Fig. 4.

Transmit unit **50** receives data **52** to be transmitted; in this case a stream of binary data. Data **52** is delivered to a transmit processing block **56**. Transmit processing block **56** subdivides data **52** into a number k of parallel streams. Then, processing block **56** applies an encoding mode to each of the k streams to thus encode data **52**. It should be noted, that before transmission data **52** may be interleaved and pre-coded by an interleaver and a pre-coder (not shown). The purpose of interleaving and pre-coding is to render the data more robust against errors. Both of these techniques are well-known in the art.

The mode is determined by a modulation during which data **52** is mapped into a constellation at a given modulation rate, and a coding rate at which this translation is performed. For example, data **52** can be converted into symbols through modulation in a constellation selected from among PSK, QAM, GMSK, FSK, PAM, PPM, CAP, CPM or other suitable constellations. The transmission rate or throughput of data **52** will vary depending on the modulation and coding rates used in each of the k data streams.

Table 1

Mode	Modulation Rate (bits/symbol)	Coding Rate	Throughput (bits/s/Hz)
1	2	3/4	3/2
2	2	2/3	4/3
3	2	1/2	1
4	2	1/3	2/3

5	4	3/4	3
6	4	2/3	8/3
7	4	1/2	2
8	4	1/3	4/3
9	5	3/4	15/4
10	5	2/3	10/3
11	5	1/2	5/2
12	5	1/3	5/3
13	6	3/4	9/2
14	6	2/3	4
15	6	1/2	3
16	6	1/3	2

Table 1 illustrates some typical modes with their modulation rates and coding rates and the corresponding throughputs for data **52**. The modes are indexed by a mode number so as to conveniently identify the modulation and coding rates which are to be applied to data **52** in each mode. Lookup tables analogous to Table 1 for other coding and modulation rates can be easily derived as these techniques are well-known in the art.

Referring back to Fig. 3, a set of modes, arranged conveniently in the form of lookup table indexed as described above, is stored in a database **78** of transmit unit **50**. Database **78** is connected to a controller **66**, which is also connected to transmit processing block **56** and spatial mapping unit **58**. Controller **66** controls which mode from database **78** is applied to each of the k streams and spatial mapping to be performed by spatial mapping unit **58**.

In addition to encoding the k streams, transmit processing block **56** adds training information into training tones **T** (see Fig. 5) and any other control information, as is known in the art. Thus processed, the k streams are sent to an up-conversion and RF amplification stage **70** having individual digital-to-analog converters and up-conversion/RF amplification

blocks **74** through the spatial mapping unit **58.** The spatial mapping unit **58** maps the k streams to M inputs of the up-conversion and RF amplification stage **70.** The M outputs of amplification stage **70** lead to corresponding M transmit antennas **72** of an antenna array **76.**

A person skilled in the art will recognize that the number M of transmit antennas **72** does not have to be equal to the number of streams k. That is because various spatial mappings can be employed in assigning streams k to transmit antennas **72.** In one mapping, a certain transmit antenna **72B** transmits one of the k streams. In another mapping, a number of transmit antennas **72** transmit the same stream k. In yet another embodiment, the k streams are assigned to M antennas **72** or a subset thereof via the spatial mapping unit **58** and the unit **70.** In fact, any kind of mapping involving the use of spatial multiplexing (SM) and antenna diversity can be used in the method and system of the invention.

Transmit antennas **72** transmit data **52** in the form of transmit signals TS. Fig. 5 illustrates, as will be recognized by those skilled in the art, a multicarrier transmission scheme with n frequency carriers (tones). The vertical axis illustrates frequency carriers while the horizontal axis illustrates OFDM symbol periods. Each block corresponds to one of n frequency carriers during an OFDM symbol. The blocks marked with D correspond to data and the blocks marked with T correspond to training.

Fig. 5 indicates that training is performed on all tones during an OFDM training symbol, it will be clear to a person skilled in the art that a subset of these tones could be used for

training and the corresponding frequency response could be computed at the receiver by interpolating.

Transmit signals TS propagate through channel **22** and there experience the effects of changing conditions of channel **22**, as described above. Transmit signals TS are received in the form of receive signals RS by a receive antenna **91A** belonging to an antenna array **92** of a receive unit **90**, shown in Fig. 4.

Again referring to Fig. 4, receive unit **90** has N receive antennas **91A, 91B, ..., 91N** for receiving receive signals RS from transmit unit **50**. Receive unit **90** can be any suitable receiver capable of receiving receive signals RS via the N receive antennas **92**. Exemplary receivers include linear equalizer receivers, decision feedback equalizer receivers, successive cancellation receivers, and maximum likelihood receivers.

Receive unit **90** has an RF amplification and down-conversion stage **94** having individual RF amplification/down-conversion/ and analog-to-digital converter blocks **96** associated with each of the N receive antennas **91A, 91B, ..., 91N**. The N outputs of stage **94** are connected to a receive processing block **98** which performs receive processing to recover the k streams encoded by transmit processing block **56** of transmit unit **50**. The recovered k streams are passed on to a signal detection, decoding and demultiplexing block **100** for recovering data **52**. In the case of antenna diversity processing it should be understood that k is equal to one thus there is only a single stream recovered.

The receive processing block **98** computes the quality parameters for each of k streams and sends this information to a

statistics computation block for computing statistical parameters of the one or more quality parameters. The method of the invention can recognize slow and rapid channel variations and allows for efficient mode selection by taking both types of variations into account. This is accomplished by taking into account at least two statistics of one or more quality parameters. This may include either or both short-term and long-term quality parameters. Suitable short-term quality parameters include signal-to-interference and noise ratio (SINR), signal-to-noise ratio (SNR) and power level. Suitable long-term quality parameters include error rates such as bit error rate (BER) and packet error rate (PER).

For example, in one embodiment, the first-order and second-order statistics are derived from a short-term quality parameter such as the SINR. In another embodiment statistics of both a short-term and a long-term quality parameter are used.

In the present embodiment the short-term quality parameter used is SINR. Statistics computation block **102** computes a first-order statistical parameter **104** and a second-order statistical parameter **106** of SINR. Conveniently, first-order statistical parameter **104** is mean SINR and second-order statistical parameter is a variance SINR. Variance **106** of SINR actually consists of two values, SINR temporal variance **106A** and SINR frequency variance **106B.** In systems which do not employ multi-carrier transmission schemes frequency variance **106B** does not have to be computed. It should be noted that each data stream of k streams will have an associated statistical parameter **104,106A, 106B.**

A window adjustment **108** such as a timing circuit is connected to statistics computation block **102.** Window adjustment **108**

sets a first time interval or first sampling window τ_1 (see Fig. 5) during which the SINR is sampled. Conveniently, SINR is sampled during training tones T occurring during sampling window τ_1. The present embodiment uses multiple carrier frequencies f_c and thus the SINR is sampled and computed by block **102** for data **52** transmitted at each of the n carrier frequencies f_c. By buffering the SINR values for all the training tones T during time window τ_1 statistics computation block **102** constructs the following matrix:

$$\begin{bmatrix} SINR_{1,1} & SINR_{1,2} & \cdots & SINR_{1,w} \\ SINR_{2,1} & \cdots & & \\ \cdots & & & \\ SINR_{n,1} & & & SINR_{n,w} \end{bmatrix}$$

where $SINR_{i,j}$ is the SINR at the i-th carrier frequency f_{ci} during training phase j. There are thus 1 to n carrier frequencies f_c and 1 to w training phases.

First-order statistical parameter **104** of short-term quality parameter, in this case mean SINR, can be expressed as:

$$SINR_{mean} = \frac{1}{n \cdot w} \sum_{i=1}^{n} \sum_{j=1}^{w} SINR_{i,j} \ .$$

Second-order statistical parameters **106A, 106B** of short-term quality parameter, in this case SINR frequency variance and SINR time variance can be expressed as:

$$SINR_{\mathrm{var}(freq)} = \frac{1}{n \cdot w} \sum_{i=1}^{w} \sum_{j=1}^{n} \left[SINR_{j,i} - \frac{1}{n} \sum_{k=1}^{n} SINR_{k,i} \right]^2, \text{ and}$$

$$SINR_{\mathrm{var}(time)} = \frac{1}{w}\sum_{k=1}^{w}\left[\frac{1}{n}\sum_{i=1}^{n} SINR_{i,k} - \left(SINR_{mean}\right)\right]^2$$

In general, the duration of first sampling window τ_1 takes into account general parameters of the communication system and/or channel **22**. For example, channel **22** has a coherence time during which the condition of channel **22** is stable. Of course, the coherence time will vary depending on the motion of receive unit **90**, as is known in the art. In one embodiment, window adjustment **108** sets first sampling window τ_1 based on the coherence time. Specifically, first sampling window τ_1 can be set on the order of or shorter than the coherence time. Thus, the first- and second-order statistical parameters **104**, **106A**, **106B** computed during time window τ_1 are minimally affected by loss of coherence. In another embodiment window adjustment **108** sets first sampling window τ_1 to be much larger than the coherence time.

Alternatively, window adjustment **108** sets first sampling window τ_1 on the order of or equal to a link update time or a delay time. This is the amount of time required for receive unit **90** to communicate to transmit unit **50** the SINR statistics and/or the appropriate mode selection based on these SINR statistics as explained below. The delay time is a design parameter limited by the complexity of computations involved and feedback from receive unit **90** to transmit unit **50**. Of course, the delay time of the system should preferably be less or significantly less than the channel coherence time when the window size τ_1 is chosen smaller than the coherence time. On the converse, when the delay time of the system is greater than channel coherence

time the window size τ_1 should also be chosen appropriately to be larger than the channel coherence time.

It should be noted that the first-order and second-order statistics of the short-term quality parameter, in the present case mean and variance of SINR could be sampled and computed over different sampling windows.

In accordance with yet another alternative, SINR frequency variance **106A** and SINR time variance **106B** require a variance computation time. Variance computation time is chosen as the minimum amount of time required to obtain an accurate value of variances **106A**, **106B**. Window adjustment **108** therefore sets first sampling window τ_1' on the order of or equal to the variance computation time, the embodiment illustrated in Fig. 5 shows τ_1 and τ_1' to be equal.

In addition to the above computations, receive unit **90** also computes a long-term quality parameter, in this embodiment an error rate of data **52**. For example, receive unit **90** computes a bit error rate (BER) or a packet error rate (PER) of data **52**. It typically takes a much longer period of time than the length of first sampling windows τ_1, τ_1' to compute these error rates. Conveniently, these error rates are thus computed during second sampling window τ_2 or over an error rate computation time (see Fig. 5). The computation of these error values and circuits necessary to perform these computations are well known in the art.

It should be noted that long-term quality parameters can be sampled over two second sampling window lengths as well. In the present embodiment only first-order statistical parameter, mean BER is computed during second time window τ_2.

In the present embodiment long-term quality parameter computed is the packet error rate (PER). As is well known in the art, the packet error rate can be computed by keeping track of the cyclic redundancy check (CRC)failures on the received packets. PER computation is a well-known technique and is performed in this embodiment by a PER statistics circuit **110**. The PER computation can be used to further improve mode selection.

The first and second-order statistical parameters of the short term quality parameter **104, 106A, 106B** are delivered from statistics computation block **102** to a mode selection block **112.** The first-order statistical parameter of the long-term quality parameter, in this embodiment the mean PER is also delivered to block **112.** When used, the PER statistics circuit **110** is also connected to mode selection block **112** and delivers the PER statistics to it.

Mode selection block is connected to a database **114,** conveniently containing the same set of modes as database **78** of transmit unit **50.** The set of modes in database **114** is related to first-order statistical parameter **104** and second-order statistical parameters **106A, 106B** of short-term quality parameter.

Block **112** selects the subsequent mode number for encoding data **52.** Block **112** is connected to a feedback block **116** and a corresponding transmitter **118** for transmission of the feedback

to transmit unit **50.** Here the convenience of indexing modes becomes clear, since feedback of an index number to transmit unit **50** does not require much bandwidth. It should be noted, that in the present embodiment a mode selection is made for each of the k streams. In other words, a mode index indicating the mode to be used for each of the k streams is fed back to transmit unit **50.** In another embodiment it may be appropriate to send a mode difference indicating how to modify the current mode for subsequent transmission. For example if the current transmission is mode 1, the mode index of the subsequent mode is 3, the mode difference would be 2. In yet another embodiment, it may be suitable to send the channel characteristics back to the transmitter. In this case the computation of statistics of the quality parameter, the mode selection are performed at the transmitter.

Referring back to Fig. 3, transmit unit **50** receives feedback from receive unit **90** via a feedback extractor **80.** Feedback extractor **80** detects the mode index or any other designation of the selected modes for each of the k streams and forwards this information to controller **66.** Controller **66** looks up the mode by mode index in database **78** and thus determines the modulation, coding rate and any other parameters to be used for each of the k streams. In the event of using time-division duplexing (TDD), which is a technique known in the art, the quality parameters can be extracted during the reverse transmission from receive unit **90** or remote subscriber unit and no dedicated feedback is required.

In one embodiment, when the system of invention is placed into operation, transmit processing block **56** first assigns an initial mode, e.g., one of the modes available in the set of modes stored in database **78** to each of the k streams. The

choice of initial modes can be made on previously known data, simulations or empirical results. Transmit unit **50** then transmits data **52** in the form of transmit signals TS to receive unit **90**.

Receive unit **90** receives receive signals RS, reconstructs data **52** therefrom, and computes first-order and second-order statistical parameters **104, 106A, 106B** of short-term quality parameter. Mode selection block **112** then selects from database **114** the subsequent mode based on parameters **104, 106A, 106B.**

Table 2

SINR variance (frequency)	SINR variance (temporal)	Table
		A
		B
		C
		D

Table C

SINR mean	Mode No.
40 dB	
35 dB	
25 dB	
10 dB	

Lookup tables 2 and C illustrate a portion of database **114** arranged to conveniently determine the mode number of a subsequent mode to be used in encoding data **52** based on the frequency and temporal variances of SINR (second-order statistical parameters of short-term quality parameter) and

mean SINR (first-order statistical parameter of short-term quality parameter). Table 2 is referenced to additional tables A, B, C and D (only table C shown) based on frequency and temporal variances **106A, 106B** of SINR. For example, the third entry in Table 2 corresponds to table C where modes are ordered by mean SINR (first-order statistical parameter of short-term quality parameter). Thus, a subsequent mode to be applied in encoding data **52** can be easily obtained from database **114** by block **112** based on its mode number.

In addition to considering the short-term quality parameters, block **112**, as an option, can obtain at least one long-term quality parameter and its statistics, i.e., first-order statistical parameter or mean PER in this embodiment. Block **112** then uses mean PER to find adjustment value δ that is added to the mean SINR value. In other words, the mean SINR value used to look up the corresponding mode is now mean SINR+δ. Thus, subsequent mode selection is altered or adjusted by taking into account the long-term quality parameter.

Once mode selection block **112** determines which modes should be used for each of the k streams, these subsequent modes are fed back to transmit unit **50** and applied to the k streams. This operation repeats itself, and each new selection of subsequent modes is fed back to transmit unit **50** to thus account for the changing conditions of channel **22**.

In fact, any combination of short-term and long-term quality parameters and their first- and second-order statistics can be used to thus appropriately select modes which should be used in transmitting data **52**. The quality parameters can further be related to link quality parameters or communication parameters

such as BER, PER, data capacity, signal quality, spectral efficiency or throughput and any other parameters to support requisite user services (e.g., voice communication). It should be noted, that BER and PER are both quality parameters and communication parameters. The subsequent mode selection can be made to optimize any of these communication parameters.

The use of at least two statistical parameters provides improved channel characterization leading to better selection of a subsequent mode than in prior art systems. Adapting the timing window to the mode update delay further improves mode selection. Furthermore, the use of long-term statistical parameters in mode selection further refines subsequent mode selection over longer time periods.

A person skilled in the art will recognize that the method and system of invention can be used in with any data transmission technique such as OFDMA, FDMA, CDMA, TDMA. It will also be clear to one skilled in the art that the above embodiments may be altered in many ways without departing from the scope of the invention. Accordingly, the scope of the invention should be determined by the following claims and their legal equivalents.

CLAIMS

What is claimed is:

1. A method for selecting a mode for encoding data for transmission in a wireless communication channel between a transmit unit and a receive unit, said method comprising:
   a) transmitting said data encoded in an initial mode from said transmit unit to said receive unit;
   b) sampling at least one quality parameter of said data received by said receive unit;
   c) computing a first-order statistical parameter of said at least one quality parameter;
   d) computing a second-order statistical parameter of said at least one quality parameter; and
   e) selecting a subsequent mode for encoding said data based on said first-order statistical parameter and said second-order statistical parameter.

2. The method of claim 1, wherein said at least one quality parameter comprises a short-term quality parameter and said method further comprises setting a first sampling window during which said short-term quality parameter is sampled.

3. The method of claim 2, wherein said wireless communication channel has a coherence time and said first sampling window is set based on said coherence time.

4. The method of claim 2, wherein said subsequent mode is applied after a delay time, and said first sampling window is set based on said delay time.

5. The method of claim 2, wherein said second-order statistical parameter is a variance of said short-term quality parameter, said variance being computed over a variance computation time and said first sampling window being set on the order of said variance computation time.

6. The method of claim 2, wherein said short-term quality parameter is selected from the group consisting of signal-to-interference and noise ratio, signal-to-noise ratio and power level.

7. The method of claim 1, wherein said at least one quality parameter comprises a long-term quality parameter and said method further comprises setting a second sampling window during which said long-term quality parameter is sampled.

8. The method of claim 7, wherein said first-order statistical parameter is a mean of said long-term quality parameter, said mean being computed over a mean computation time and said second sampling window being set on the order of said mean computation time.

9. The method of claim 7, wherein said long-term quality parameter comprises an error rate of said data at said receive unit.

10. The method of claim 9, wherein said error rate is computed over an error rate

computation time, and said second sampling window is set on the order of said error rate computation time.

11. The method of claim 9, wherein said error rate is selected from the group consisting of bit error rate and packet error rate.

12. The method of claim 1, wherein said first-order statistical parameter is a mean of said at least one quality parameter.

13. The method of claim 1, wherein said second-order statistical parameter is a variance of said at least one quality parameter.

14. The method of claim 13, wherein said data is transmitted at more than one frequency and said variance is a frequency variance.

15. The method of claim 13, wherein said data is transmitted in a multi-carrier scheme and said variance is a frequency variance.

16. The method of claim 13, wherein said variance is a temporal variance.

17. The method of claim 1, wherein said initial mode is selected from a set of modes related to said at least one quality parameter.

18. The method of claim 1, wherein said subsequent mode is selected to maximize a communication parameter.

19. The method of claim 18, wherein said communication parameter is selected from the group consisting of bit error rate, packet error rate, data capacity, signal quality, spectral efficiency and throughput.

20. The method of claim 1, further comprising communicating said subsequent mode to said transmit unit.

21. The method of claim 1, wherein at least one of said transmit unit and said receive unit are multiple input and multiple output units.

22. The method of claim 1, wherein said transmitting step is performed in accordance with a transmission technique selected from the group consisting of OFDMA, FDMA, CDMA, TDMA.

23. A method for selecting a mode from a set of modes for encoding data for transmission in a wireless communication channel between a transmit unit and a receive unit, said method comprising:

a) transmitting said data encoded in an initial mode selected from said set of modes from said transmit unit to said receive unit;

b) sampling a short-term quality parameter of said data received by said receive unit;

c) computing a statistical parameter of said short-term quality parameter;

d) selecting a subsequent mode from said set of modes for encoding said data based on said short-term statistical parameter;

e) sampling a long-term quality parameter of said data received by said receive unit; and

f) adjusting said subsequent mode selection in step d) based on said long-term quality parameter.

24. The method of claim 23, wherein said long-term quality parameter is an error rate selected from the group consisting of bit error rate and packet error rate.

25. The method of claim 23, further comprising setting a first sampling window during which said short-term quality parameter is sampled and setting a second sampling window during which said long-term quality parameter is sampled.

26. The method of claim 25, wherein said long-term quality parameter is an error rate and is computed over an error rate computation time, and said second sampling window is set on the order of said error rate computation time.

27. The method of claim 25, wherein said short-term quality parameter is selected from the group consisting of signal-to-interference and noise ratio, signal-to-noise ratio and power level.

28. The method of claim 23, wherein said set of modes is arranged in a lookup table and ordered by said short-term quality parameter.

29. The method of claim 28, wherein said adjusting comprises modifying said lookup table based on said long-term quality parameter.

30. The method of claim 23, wherein said statistical parameter comprises at least one statistical parameter selected from the group consisting of first-order statistical parameters and second-order statistical parameters.

31. A system for assigning a subsequent mode for encoding data for transmission in a wireless communication channel, said system comprising:

a) a transmit unit having a transmit processing block for encoding said data in a mode;

b) a receive unit for receiving said data transmitted from said transmit unit, said receive unit having:

1) a statistics computation block for sampling at least one quality parameter of said data and computing a first-order statistical parameter of said at least one quality parameter and a second-order statistical parameter of said at least one quality parameter;

2) a mode selection block for assigning said subsequent mode based on said first-order statistical parameter and said second-order statistical parameter.

32. The system of claim 31, further comprising at least one database containing a set of modes from which said mode and said subsequent mode are selected.

33. The system of claim 32, wherein said at least one database comprises a first database in said transmit unit and a second database in said receive unit.

34. The system of claim 31, further comprising a feedback mechanism for communicating said subsequent mode between said receive unit and said transmit unit.

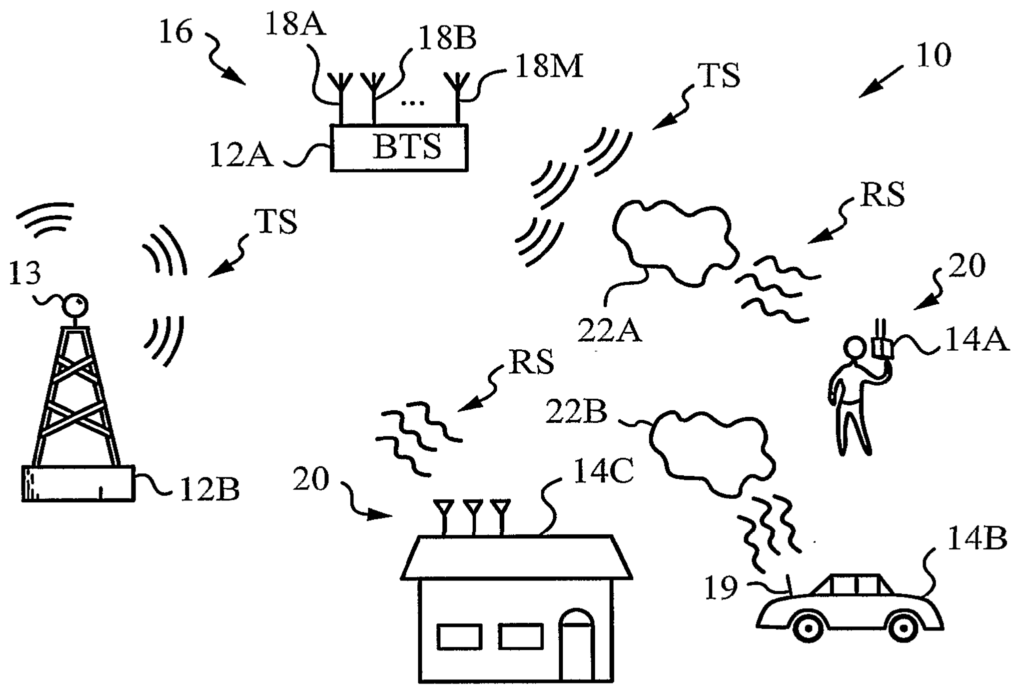

16
18A
18B
18M
10
TS
12A
BTS
RS
TS
20
13
22A
14A
RS
22B
20
14C
12B
14B
19

FIG. 1

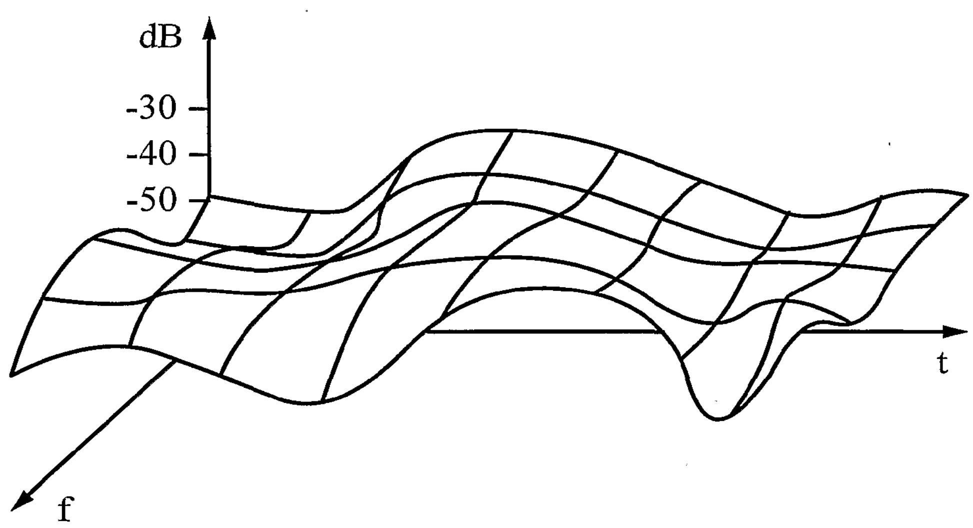

dB
-30
-40
-50
t
f

FIG. 2

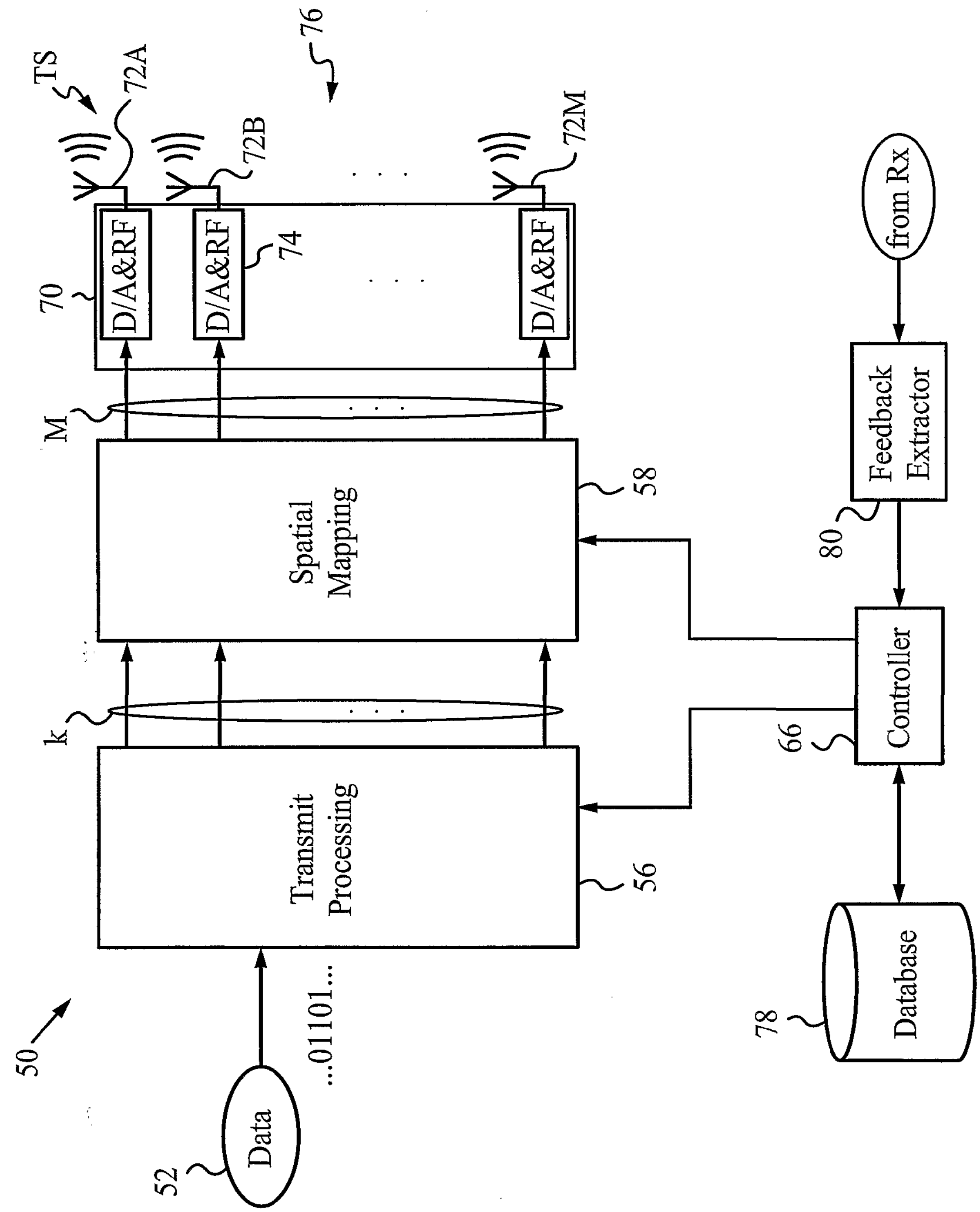

TS
72A
76
72B
72M
70
D/A&RF
74
D/A&RF
D/A&RF
from Rx
M
Feedback Extractor
Spatial Mapping
58
80
k
Controller
66
Transmit Processing
56
50
...01101...
Database
78
Data
52

FIG. 3

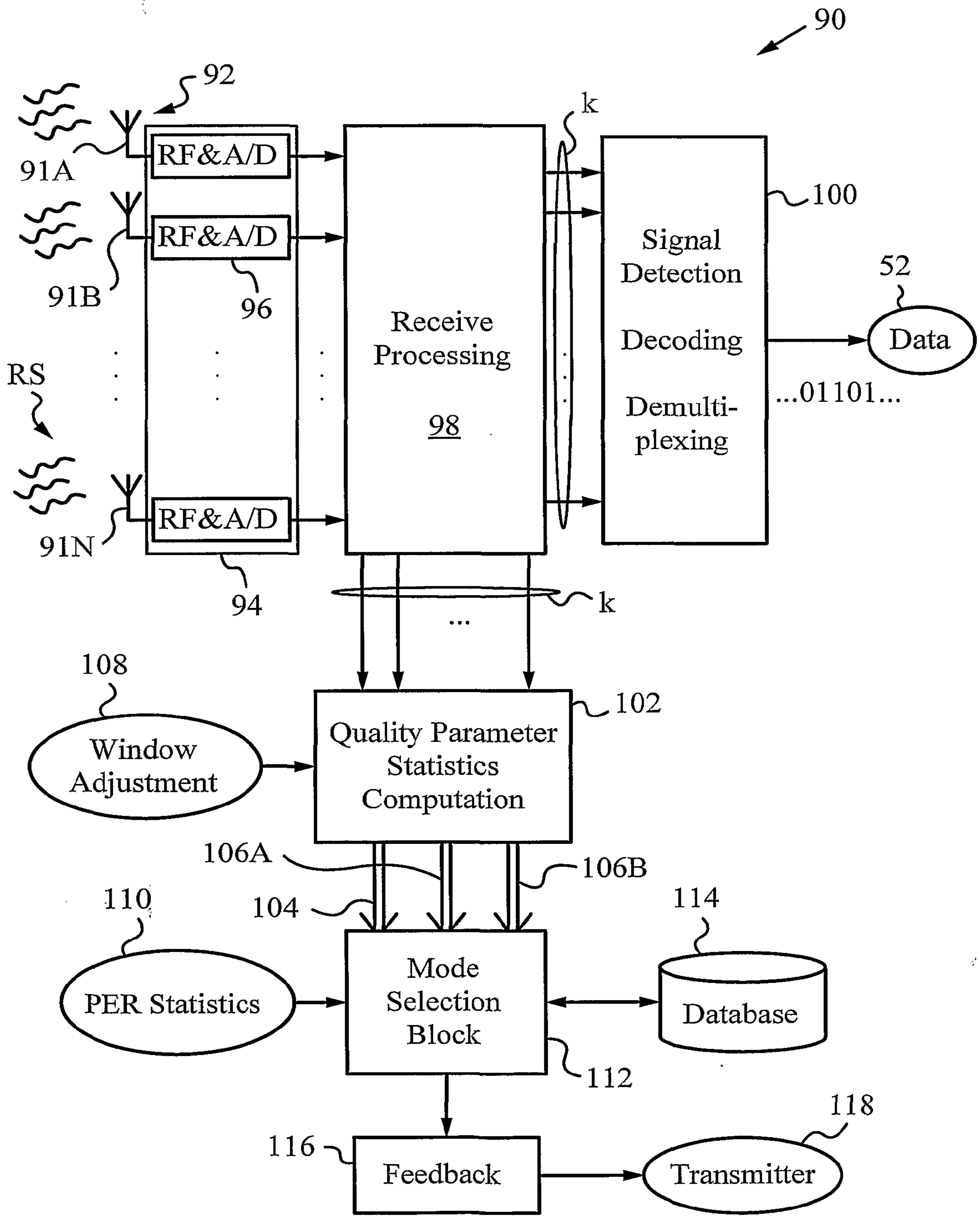

90
92
91A
91B
91N
RS
RF&A/D
RF&A/D
RF&A/D
96
94
Receive Processing
98
k
Signal Detection
Decoding
Demulti-plexing
100
52
Data
...01101...
k
...
108
Window Adjustment
Quality Parameter Statistics Computation
102
106A
106B
104
110
PER Statistics
Mode Selection Block
114
Database
112
116
Feedback
118
Transmitter

FIG. 4

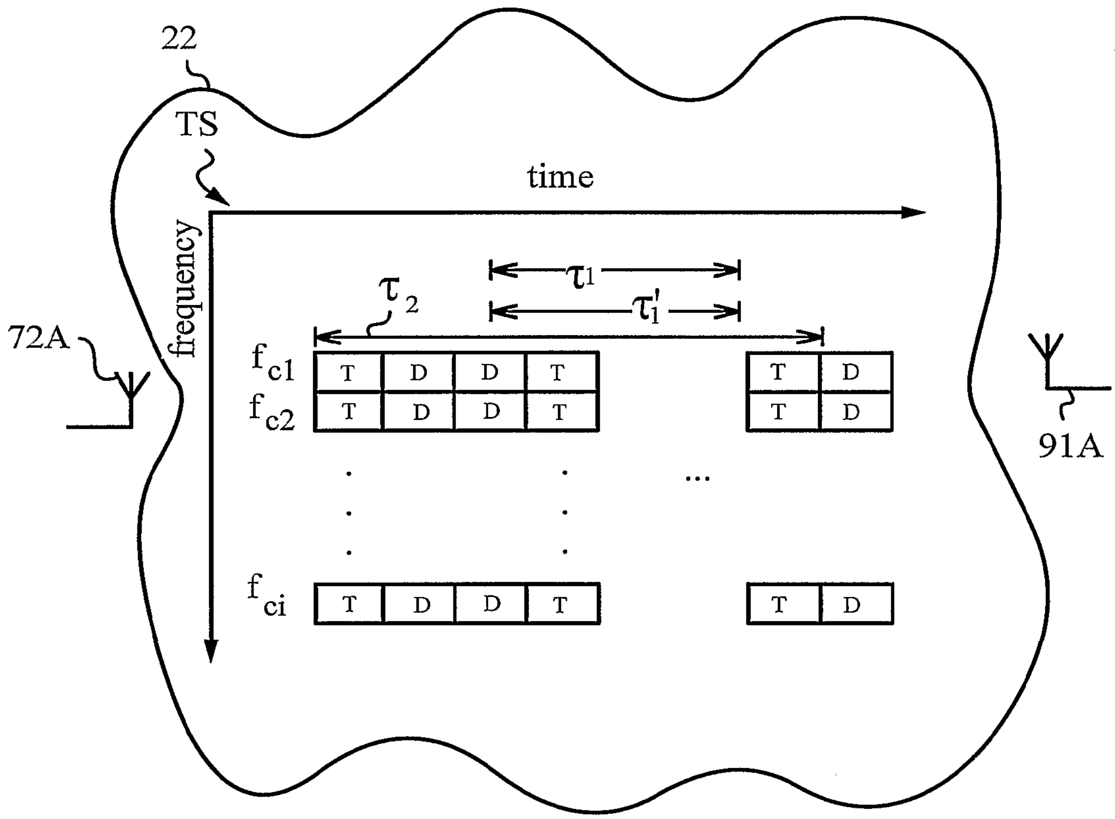


FIG. 5